ELECTRICAL COMMUNICATION

Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies

ALEXANDER GRAHAM BELL—SCIENTIST
FRANCE-ENGLAND SUBMARINE CABLE AND PARIS-CALAIS CABLE
YARN-NUMBERING SYSTEM
BEHAVIOUR OF TELEPHONE EXCHANGE TRAFFIC
TECHNICAL COORDINATION ON AN INTERNATIONAL BASIS
EXTENSION OF NORTON'S METHOD OF IMPEDANCE TRANSFORMATION
BIKINI OBSERVATIONS AND THEIR SIGNIFICANCE
LANDING AIRCRAFT WITH GROUND RADAR
FREQUENCY-MODULATED BROADCAST TRANSMITTERS
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TELEPHONE STATISTICS OF THE WORLD
C.C.I.F. HONORS TWO PIONEERS
RECENT TELECOMMUNICATION DEVELOPMENTS

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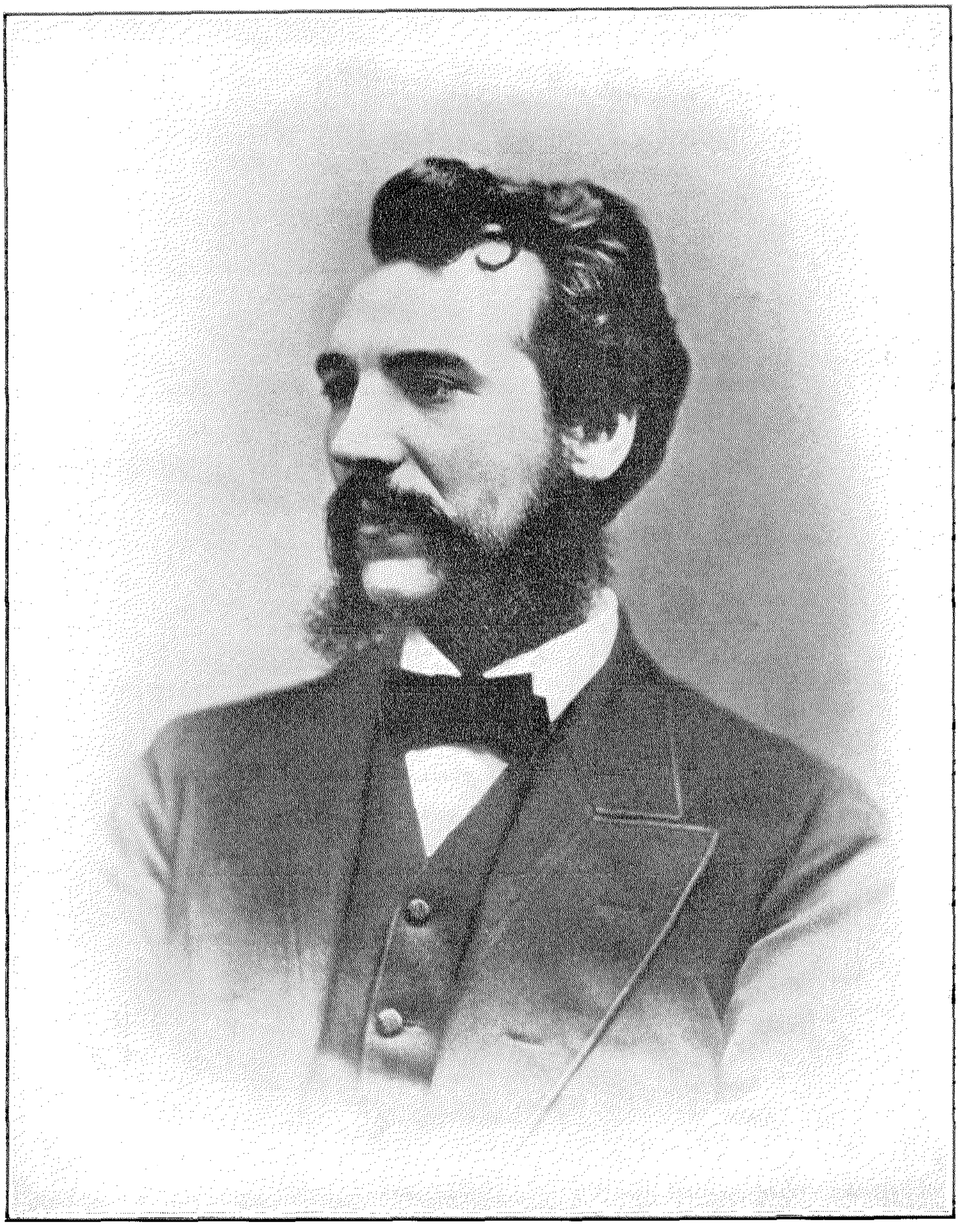

Alexander Graham Bell as he appeared in 1876, the year the telephone patent was granted.

Alexander Graham Bell—Scientist*

By F. J. MANN

Managing Editor, Electrical Communication

ALEXANDER GRAHAM BELL was born one hundred years ago, March 3, 1847, in Edinburgh, Scotland. In commemorating the centennial of his birth, he will be eulogized as a philanthropist, a teacher, and as a man who did much to aid the deaf and hard of hearing, as well as the inventor of the telephone. These he was, it is true. But, judging from his own life and writings, he would have preferred to be remembered primarily as a scientist.

By present standards, Bell might not have qualified as a great scientist. It must be remembered, though, that he was born into a world only slowly awakening to the wonders of scientific discovery. Only a few years before his birth, the Morse telegraph (1832), the McCormick reaper (1833), photography (1839), ether used as an anaesthetic (1842), and Howe's sewing machine (1845) had come into being. Compared with research on atomic fission and development of radar, these inventions were the results of crude efforts often carried on over relatively long periods of time in attics or small sheds. But, like Bell's invention of the telephone, they had something in common. They were basic accomplishments, the results of creative thought, often of a single mind.

Nothing in history indicates that these achievements, and many others that followed, rose full-blown from flash-of-genius ideas. Each involved its own painstaking effort. Often never recorded, were the heartbreaking stories of failure by those who struggled toward common goals, but lacked the creative spark essential to success. Bell, before he invented the telephone, overcame numerous difficulties. He had his contemporary rivals, who strove for the same objective—to send the human voice over wires. Credit Bell's triumph to a singularly inventive mind, if you will, or to a "lucky break," as some thought. It was Bell's belief, as well as that of many of his colleagues, that his success was due to his unique training and lifelong association with the science of speech and acoustics.

Bell was born into a family preoccupied with the correction and teaching of speech. Bell's grandfather, Alexander Bell (1790–1865), who started life as a shoemaker in St. Andrews, Scotland, where for generations his ancestors had been shoemakers, forsook this occupation for that of a Shakespearean actor. After he had married and moved to Edinburgh, he abandoned the stage to become an elocutionist and a "corrector of defective utterance." The grandfather, then, was the first in the family to study the mechanism of speech with the object of correcting defects by explaining to his pupils the correct positions of the vocal organs in uttering sounds.

This profession, founded by the grandfather, became a family profession which was carried on by his children and grandchildren. Both of his sons, for example, David Charles Bell (1817–1902) and Melville Bell (1819–1905), were elocutionists and correctors of defective speech. It should be pointed out, though, that neither the grandfather nor his sons could be called mere reciters or practitioners of cures for defects of speech. They studied the anatomy of speech with scientific thoroughness. Among them they are said to have exerted a strong influence on English speech.

Melville Bell, Alexander Graham Bell's father, won a world-wide reputation in his profession. His *Standard Elocutionist* had run through 168 editions by 1893 and it has sold impressively since then. By the end of his life, some 26 of his textbooks and charts on speech and phonetics were in use in schools and colleges.

Bell's Father Invents Visible Speech

When Alexander Graham Bell was in his early teens, his father branched off from the profession of teaching into that of inventor. He called his creation "visible speech," a remarkable system

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* Reprinted from *Electrical Engineering*, v. 66, n. 3; March, 1947.

of symbols for depicting the actions of the vocal organs in uttering sounds. These symbols could be used in printed form like letters of the alphabet. He claimed that what he had really devised was a universal single alphabet capable of expressing the sounds of all languages and that his letters, instead of being arbitrary characters, were symbolic representations of the organs of speech and their positions in uttering sounds. Visible speech was actually a code of symbols made up of curved and straight lines—resembling shorthand—that indicated the position and action of the throat, tongue, and lips in pronouncing syllables or various sounds.

Visible speech proved useful as a key to the pronounciation of words in all languages. It also developed that the symbols could be used to guide the deaf in learning to speak. Alexander Graham Bell, as he grew up, became expert in the use of visible speech, particularly in aiding deaf pupils, a fact which had a most important bearing on his life after he came to America.

Early Interest in Music

Despite this environment of speech correction and scientific investigation, it seemed that young Alexander would follow the arts instead of the family profession. As a boy, Bell exhibited an outstanding aptitude for the piano. "Music especially was my earliest hobby," said Bell. He learned to play at such an early age that, in later life, he was unable to recall a time when he could *not* play. He seemed to have picked it up by himself without any instruction, and, although he knew nothing of written music, he could play anything he heard and could improvise at length. This acute ear for music was destined one day to pick up the faint "ping" of a reed accidentally plucked in a Boston attic and to recognize in it the possibility of electrical transmission of speech.

Signor Auguste Benoit Bertini, a distinguished professor of music, heard young Bell play and offered to give him instruction in his system of reading music by sight. Bell worked hard at his lessons and became such a promising pupil that Bertini wanted him to become his successor. However, Bertini died before this hope might have been realized. After Bertini's death, Bell received no further formal instruction in music except from his mother who sought to carry out Bertini's ideas as well as she could. But the old professor seems to have inspired young Bell with a passion for music, which led him, for a time, to consider a musical profession. In fact, he did teach music for two years at Weston House, a boy's school in Elgin, Scotland, starting in 1863 when he was only 16 years old.

Urged to Follow Family Profession

When we consider today how the same amplifier and loudspeaker reproduce both speech and music, a musical career and the teaching of speech correction are evidently more nearly related than was then apparent. At any rate, Bell's father urged him to give up music as a profession and carry on the family practice in the art and science of speech.

Aleck, as the family called him, had good reason to follow his father's advice. Not only had the father achieved distinction in his profession, but he seems to have been singularly successful in what is often a much more difficult task, that of instilling in his sons a sense of intellectual curiosity and integrity as well as individuality of thought. His influence was exerted in various ways. One was to encourage his boys to make collections of all sorts and to arrange the specimens in accordance with their own ideas rather than in conformity with those of others. Aleck's collection consisted of a large number of skeletons, neatly arranged and classified as in a museum. It included a goodly number of skulls of squirrels, rabbits, cats, and dogs, but the gem of the collection was a real human skull presented to him by his father.

Alexander Graham Bell put great store in this phase of his education. In the last year of his life he wrote:[1]

> I can see in these natural-history collections a preparation for scientific work. The collection of material involved the close observation of the liknesses and differences of objects of very similar kind, and the orderly arrangement, as in a museum, stimulated the formation of generalizations of various kinds. . . . I am inclined to think that the making of these collections formed an important part of my education and was responsible for my early bent toward scientific pursuits.

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[1] A. G. Bell, "Prehistoric Telephone Days," *The National Geographic Magazine*, v. 16, pp. 226–227; March, 1922.

Indifferent to Classical Studies

While Aleck excelled in these exceptional pursuits, he was not considered brilliant at his formal studies. From 1855 to 1859, he attended McLaren's Academy and the Royal High School in Edinburgh, but he lacked enthusiasm for the prescribed subjects and especially disliked Latin and Greek, then considered the mainstays of formal education. His opinion of the advantage of scientific over classical training as a preparation for life was expressed in his prophetic address to the graduates of McKinley Manual Training School at Washington, D. C., in 1917:

> The nation that fosters science becomes so powerful that other nations must, if only in self-defense, adopt the same plan. . . . Scientific and technical experts are destined in the future to occupy distinguished and honorable positions in all the countries of the world.

It is to his grandfather that a great deal of the credit must go for inspiring and encouraging Aleck to overcome his indifference to formal education. In 1860, when 13, Aleck spent a year with his grandfather in London. His grandfather helped him map out his time and devote certain hours to the ordinary school subjects. He also gave him personal lessons in elocution and English literature. At the same time, Aleck received instruction in the mechanism of speech from his grandfather, who permitted the boy to be present at the instruction of some of his pupils so that he might observe the methods of correcting defective utterance.

This year with his grandfather converted Aleck from what he described as "an ignorant and careless boy into a rather studious youth, anxious to improve his educational standing by his own exertions and fit himself for college." The conversion was really from indifference to enthusiasm, for throughout every stage of his development and education, even in later years, his enthusiasm governed his working habits. In fact, in the subjects he loved, like music and science, he was almost too enthusiastic so that, for many years, he was constantly threatened with breakdowns. However, rather than an over-studious, bookish person, he was fun-loving and ardent, so gifted that he took vast pleasure in whatever happened to interest him.

Bordering on the precocious was an incident Bell often recalled; it led to what he considered to be his first invention and was related by him in 1922:[2]

> When I was quite a little fellow, it so happened that my father had a pupil of about my own age with whom I used to play. He was the son of a Mr. Herdman, who owned large flour mills near Edinburgh, and, of course, I went over to the mills pretty often to play with him there. We romped about and got into all sorts of mischief, until at last one day Mr. Herdman called us into his office for a very serious talk.
>
> "Why can't you boys do something useful," he said, "instead of always getting into mischief?"
>
> I mildly asked him to tell us some useful thing to do, and he replied by putting his arm into a bag and pulling out a handful of wheat. He showed us that the grains were covered with husks, and said: "If you could only take the husks off that wheat, you'd be doing something useful indeed."
>
> That made rather an impression upon my mind, and I began to think, "Why couldn't we take the husks off *by brushing the seeds with a nailbrush?*"
>
> We tried the experiment and found it successful, although it involved a good deal of hard work from the two mischief-makers. We persevered, however, and soon had a nice little sample of cleaned wheat to show Mr. Herdman. I then remembered that during our explorations at the mills we had come across a large vat or tank with a paddle-wheel arrangement in it that whirled round and round in a casing of quite rough material, brushes of fine wire netting, or something of that sort. If we could only put the wheat into the machine, I thought, the whirling of the paddle should cause the seeds to rub against the rough surface of the casing, and thus brush off the husks.
>
> It was a proud day for us when we boys marched into Mr. Herdman's office, presented him with our sample of cleaned wheat, and suggested paddling wheat in the dried-out vat.
>
> "Why," said Mr. Herdman, "that's quite a good idea," and he immediately ordered the experiment to be made. It was successful, and the process, I understand, or a substantially similar one, has been carried on at the mills ever since.

Meets Well-Known Scientists

If this first "invention" had nothing to do with the telephone, later experiments made by Bell when he was 15 years old most certainly did. The father's fame as a scientist of speech problems brought him in contact with some of the most outstanding men of his day. His evident pride in his sons and his interest in their development led him to share these valuable and pleasant associations with them. Thus it was that Aleck, when still a boy, met such men as Alexander J. Ellis, the translator of Helmholtz; Max Muller, Sanskrit scholar; Henry Sweet, phonetician; Dr. Furnival, secretary of the Philological Society of London; Dr. Murray, afterwards Sir James Murray, editor of the great Oxford Dictionary; Prince Lucian Bonaparte, student of Scottish dialects; and Sir Charles Wheatstone.

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[2] Reference 1, pp. 239–241.

Visits Wheatstone

Each meeting made a vivid impression on Aleck. Probably the one to have the most immediate effect was his first contact with Sir Charles Wheatstone, who had constructed what was then called an automaton speaking machine from a description in a book on the mechanism of human speech by Baron von Kempelen. Aleck's father took him to visit Sir Charles to see the machine and hear it talk.

Aleck saw Sir Charles manipulate the machine and he heard it speak. Although the articulation was crude, it made a great impression on him. Sir Charles loaned Baron von Kempelen's book to the Bells and Aleck devoured it when he reached home. The book was in French, but he knew the language well enough to be able, with his father's assistance, to read and enjoy it.

Stimulated by their father, Aleck and his brother, Melville, attempted to construct a speaking machine of their own. They divided up the work, Melville's special part consisting of the larynx and vocal chords, to be operated by the wind chest of a parlor organ, while Aleck undertook the mouth and tongue.

Brothers Work on Speaking Machine

Melville was quite skillful in the use of tools, while Aleck was clumsy with his hands and ineffective where tools were concerned. Aleck hit upon a plan, however, that obviated the disadvantages of this defect in a great degree: he made his models of gutta-percha wherever possible.

The father took an extraordinary interest in the proposed talking-machine and encouraged the boys in every way. Bell stated that it was not until much later in life that he realized that the father looked upon the machine as a valuable educational toy which would compel his sons to become familiar with the operation of the vocal organs. It was for this reason that the father encouraged them to copy Nature itself rather than to follow in the footsteps of von Kempelen and Wheatstone.

The boys attempted to make an exact copy of the vocal organs and work the artificial lips, tongue, and soft palate by means of levers controlled by a keyboard. Aleck started on his part of the work by making a cast from the human skull his father had given him. The mouth parts of the skull were reproduced in gutta-percha. This gave a firm foundation, consisting of the upper teeth, the upper gum, the hard palate, and the back of the pharynx, with a large hole at the top representing the rear entrance into the nasal cavities. This hole was covered by a valve, consisting of a piece of wood hinged to the palate and covered with a skin of soft rubber stuffed with cotton batting. The operating lever passed through the nasal passages beyond the nose. The lips were formed of a framework of iron wire covered with rubber stuffed with cotton batting, and rubber cheeks were provided, completely closing the mouth cavity.

It was proposed to make the tongue of wooden sections, arranged side by side like the dampers of a piano. Each section was to be elevated by its appropriate lever, and the whole tongue covered by a thin skin of rubber stuffed with cotton batting. This part of the mechanism was never actually completed, but sections of the tongue were made and experimented with.

While Aleck was thus engaged, his brother made an artificial larynx, or throat, of tin, with a flexible tube attached as a windpipe. Inside the larynx were two flat sheets of tin sloping upward toward one another, but not touching at the middle. Stretched tightly upon this structure were two sheets of rubber, the edges of which did touch at the middle.

When the windpipe was blown into, the rubber vocal chords were thrown into vibration, producing a musical sound. By varying the tension of the rubber strips and by changing the force of the breath, it would squeak like a Punch and Judy show, or produce a reed-like musical tone.

When this stage had been reached, the boys were anxious to put the throat and the mouth together to see what the effect would be. Without waiting to complete the tongue, or for the arrival of the organ bellows, they fastened the tin larynx to the gutta-percha mouth and one of them blew through the windpipe. At once the character of the sound was changed; it no longer resembled a reed instrument, but a human voice. Vowel quality also could be reproduced. It could be made to sound as though someone were singing the vowel "ah."

Talking Machine a Success

Aleck opened the rubber lips a number of times in succession while his brother blew through the windpipe. The machine responded by uttering the syllables "ma-ma-ma-ma" quite clearly and distinctly. By using only two syllables and prolonging the second, they obtained good reproduction of the word "mamma," pronounced in the British fashion, with the accent on the second syllable. The next step, of course, was to try the effect on the neighbors.

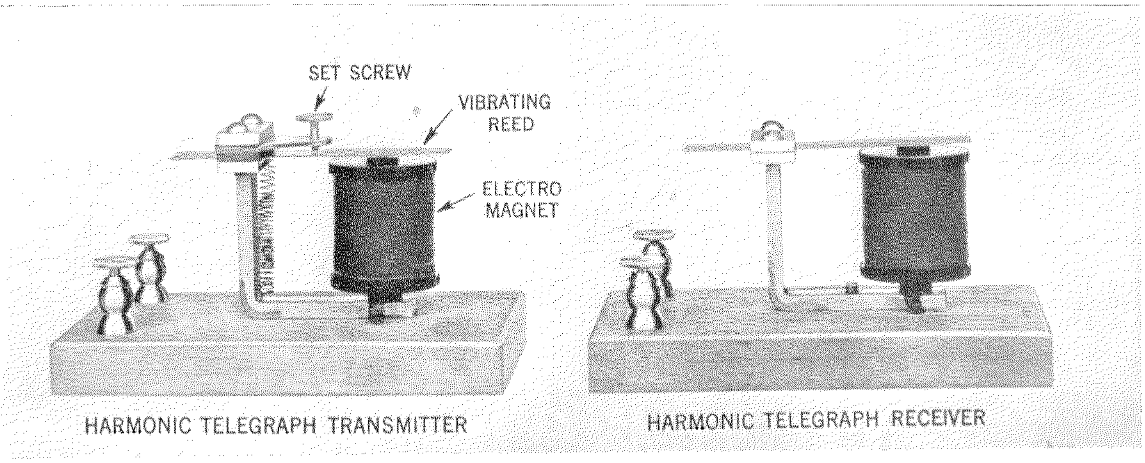
The house they lived in consisted of a number of flats that opened on a common stairway. They took the apparatus out on the landing. Melville put the windpipe to his mouth and blew for all he was worth, while Aleck manipulated the lips. The hallway resounded with loud cries of "Mamma! Mamma! Mamma!"—imitiating a child in great distress.

Soon a door opened upstairs and the boys heard a woman exclaim, "My goodness, what's the matter with that baby"!

That was all they needed. Delighted with their success, they stole quietly back into their home and gently shut the door.

Although no further work was done on the speaking machine, it had undoubtedly been successful in realizing the father's desire that through it his boys should become familiar with the structure and functions of the various vocal organs.

Aleck was always much interested in his



Replicas of the harmonic telegraph instruments used by Bell and Watson in the historic experiments of June 2, 1875. When the transmitter reed vibrated, it was supposed to touch the set-screw contact and send a pulse of current along the line. During the experiments, the reed stuck to the contact and Watson's efforts to free the reed by plucking made the current undulate. Bell, listening to the receiver, placed the free end of the reed against his ear. This made the receiver reed act like a telephone receiver diaphragm. With this arrangement, sound was actually transmitted from the transmitter to the receiver, confirming Bell's carefully worked out theories about the electrical transmission of sound.

father's examinations of the mouths of his elocution pupils. They differed in an extraordinary degree in size and shape, and yet all these variations seemed to be quite consistent with perfect speech. He then began to wonder whether there was anything in the mouth of a dog to prevent it from speaking, and started to make experiments with his intelligent Skye terrier.

Bell's Talking Dog

Stimulated by rewards of titbits, the dog was soon taught to sit up on his hind legs and growl continuously while Aleck manipulated his mouth, and to stop growling when he took his hands away. Aleck would take the dog's muzzle in his hands and open and close the jaws a number of times in succession. This resulted in the production of the syllables "ma-ma-ma" as in the case of the speaking machine.

The dog's mouth proved too small to enable Aleck to manipulate individual parts of the tongue, but by pushing upward between the bones of the lower jaw, near the throat, he found it possible to close the passageway completely at the back of the mouth. A succession of pushes of this character resulted in the syllables "ga-ga-ga."

The simple growl was an approximation of the vowel "ah" and, followed by a gradual constriction and "rounding" of the labial orifice by the hand, became converted into the diphthong "ow," as in the word "how" (ah-oo). He soon obtained the final element by itself—an imperfect "oo." The dog's repertoire of sounds finally consisted of the vowels "ah" and "oo", the diphthong "ow," and the syllables "ma" and "ga."

Dog Learns Sentence

Bell then proceeded to manufacture words and sentences composed of these elements, so that the dog could finally say the sentence "Ow-ah-oo-gamama," which, by the exercise of some imagination, readily passed for "How are you, grandmamma."

Of this talking dog, Bell wrote:[3]

> The dog soon learned that his business in life was to growl while my hands were upon his mouth, and to stop growling the moment I took them away, and we both of us became quite expert in the production of the famous sentence, "How are you, grandmamma?"
>
> The dog took quite a bread-and-butter interest in the experiments and often used to stand up on his hind legs and try to say this sentence by himself, but without manipulation was never able to do anything more than growl.
>
> The fame of the dog soon spread among my father's friends, and people came from far and near to witness the performance.

As previously stated, Bell started his teaching career at Weston House in 1863. He actually was rated as a student-teacher, that is, as compensation for teaching, he received advanced instruction at the school. A plan was worked out whereby his brother, Melville, went to the University of Edinburgh while Aleck taught music and elocution and continued his studies. At the end of the year, Bell went to the University for further classical education while Melville taught at Weston House. In 1866, he returned to Weston House as a full-time teacher. In addition, his formal education included a course in anatomy at University College, London, and he matriculated as an undergraduate at London University.

Bell Proves Vowel Sounds Compound

In 1866, while Bell was still at Weston House, an incident occurred that was destined to lead him closer to thinking about the telephone than any of his previous experiments. Recalling his efforts to make a dog talk, and how close he and his brother had come to making a talking machine, it is not surprising that he should have been curious about the sounds produced in his own mouth. Consequently, he carried on a series of experiments to determine the resonant pitches of the mouth cavities during the production of vowel sounds. He would put his mouth and tongue in position to pronounce the vowel and then tap his throat or cheek with his finger or a pencil. Since the two principal cavities are forward and back of the tongue, which moves to help form the sounds, Bell's ear could distinguish the tone of each cavity in the resonant sound that resulted from tapping in various ways. The experiments proved that vowel sounds are compound; that is, they are a combination of resonances from different vocal cavities, and modifications of a single pitch or tone.

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[3] Reference 1, p. 239.

Hears About Helmholtz Experiments

At the time Bell believed that these experiments were original. He wrote a long letter to his father explaining his "discovery about resonance pitches." His father showed the report to Alexander John Ellis, president of the London Philological Society. Ellis told Bell that the same discovery had been made by Hermann L. F. von Helmholtz and described in his classic work, *On the Sensations of Tone as a Physiological Basis for the Theory of Music.* Since Helmholtz's book had not then been translated into English (Ellis made the English translation in 1875), Ellis tried to explain the experiment to young Bell. Helmholtz, he pointed out, had not only made the same observations, but had gone further. He had built up a synthetic vowel sound from its constituents using three electrically operated tuning forks. One fork represented the pitch of the voice while the other two corresponded in pitch to the front and back cavities of the mouth in uttering a vowel sound.

Bell did not at the time have sufficient electrical knowledge to understand Ellis's description of the apparatus used by Helmholtz. He borrowed a copy of the German text of Helmholtz's book. He knew little German, but he made out just enough from the plates, or thought he did, to confuse the description Ellis had given him and to conclude that Helmholtz had sent the vowel sounds by telegraph. Bell then reasoned that if vowel sounds could be sent by telegraph, why not consonants or even speech?

Bell determined to study electricity, for he felt that it was his duty, as a student of speech, to acquaint himself with the researches of Helmholtz and to repeat Helmholtz's experiments. Consequently, 1867 found Bell, then an instructor in Somersetshire College, Bath, England, experimenting with ordinary telegraph apparatus and vainly striving to use an electromagnet to cause a tuning fork to vibrate continuously.

Three years later he learned from a French translation of the Helmholtz book that the German scientist had not telegraphed the vowel sounds. In the meantime his imagination had been stimulated by the idea. "I thought that Helmholtz had done it," Bell said later, "and that my failure was due only to my ignorance of electricity. It was a very valuable blunder. It gave me confidence. It I had been able to read German in those days, I might never have commenced my experiments!"

During this period, Bell's two brothers succumbed to tuberculosis, the younger brother, Edward Charles, in 1867 and, in 1870, the elder brother, Melville James.

Family Moves to Canada

Alexander Graham Bell also was threatened with tuberculosis. Fearing that his only remaining son might fall a victim, Bell's father resigned his lectureships, disposed of his practice in London, and emigrated to Canada. The family landed at Quebec in August, 1870, and soon settled at Tutelo Heights, near Brantford, in Ontario. The change of climate proved beneficial and Bell recovered his health in a few months.

While in Canada, Bell became interested in the dialect of the Mohawk Indians on a nearby reservation. He transcribed many of their words and phrases into visible speech. For this, the grateful Mohawks made Bell an honorary chief of North American tribes. They also taught him the war dance which he often performed thereafter in moments of triumph and which gave a number of Boston landladies some moments of anxiety.

Bell Begins Teaching in Boston

Bell went to Boston, Massachusetts, in April, 1871, to teach visible speech at the Boston School for the Deaf (now the Horace Mann School). This first assignment was temporary and lasted only two months, but other assignments followed. He became so expert in using visible speech to train deaf children to pronounce properly that in a few weeks he taught them to use more than 400 English syllables, some of which they had failed to learn in two or three years through other teaching methods. Bell's method was to teach the pupil to relate the symbol, by touch or demonstration, to the vocal process it indicated. For example, the consonant "P" calls for placing the tongue against the lower teeth, closing the lips, and then opening them to ejaculate a puff of air. After writing the symbols for these processes on a blackboard, Bell would open his own mouth to show the tongue position, then hold the pupil's hand in position to feel the puff of air. Similar procedures were repeated for other letters.

Becomes Professor at Boston University

On October 1, 1872, Bell opened his own school of vocal physiology at 35 West Newton Street, Boston, to receive as pupils deaf mutes, persons with defective speech but with normal hearing, and teachers of the deaf and dumb. The following year, he became Professor of Vocal Physiology in the School of Oratory at Boston University. Bell then conducted his classes there, and retained his position on the faculty until 1877.

It was about this time that Bell met two men who were to have a marked influence on his life and the invention of the telephone. The first was Gardiner Greene Hubbard, an attorney inclined toward promotion of civil and governmental progress. The second was Thomas Sanders, a Boston leather merchant.

Hubbard's daughter, Mabel, had been left deaf after an attack of scarlet fever when she was four years old. In his efforts to help his daughter, Hubbard was confronted with school authorities who held the opinion that the deaf had no place in normal society—that the best that could be done for them was to send them to an institution where they could be taught the sign language. Hubbard, however, refused to accept this fate for his daughter and proceeded to do everything possible to save Mabel's power of speech. He even sent her to Germany for instruction in an oral method.

When Bell arrived in Boston with his visible-speech method, Hubbard naturally became his strongest champion. As a matter of fact, Hubbard had, shortly before, succeeded in having a bill passed in the Massachusetts legislature for the establishment of an oral school for the deaf, over strong opposition based on the premise that any attempt to teach "deaf mutes" to speak "flouted the decrees of Providence." The common interest of the two men in oral-speech training for the deaf resulted in the growth of a strong bond of friendship between them.

Bell was often a welcome guest at the spacious Hubbard home in Cambridge, Massachusetts. During one of these many visits, Bell illustrated some of the mysteries of acoustics with the aid of a piano. "Do you know," he said, "that if I sing the note *G* close to the strings of the piano, that the *G* string will answer me?"

"Well, what then?" asked Hubbard.

"It is a fact of tremendous importance," replied Bell. "It is evidence that we may some day have a musical telegraph which will send as many messages over one wire as there are notes on that piano."

Hubbard could see the value of such a device and later financed Bell's experiments with the harmonic telegraph. Being a practical man of business he could see how such a device might simplify the telegraph business and reduce costs. But he could not, at the same time, see Bell's idea of sending speech over an electric wire as anything but a worthless scientific toy. Such opposition, for a long time, retarded any attempts on the part of Bell to develop a telephone. Nevertheless, Hubbard's legal as well as financial encouragement were invaluable to Bell in filing his patent claims properly, gaining recognition from important persons, and in many other ways.

Frequent Visitor at Hubbard Residence

Not by any means the least important advantage to Bell of Hubbard's friendship was the opportunity to visit at the Hubbard residence. Bell was often invited for Sunday dinner which, at that busy time of his life, was just about the only social recreation and relaxation he had. On these visits he came to know the four Hubbard daughters and, in time, fell in love with Mabel. They were married July 11, 1877.

Bell's friendship with Thomas Sanders also started through his work of teaching the deaf to speak. For Sanders had a son, George, who had been born deaf. George Sanders became a private pupil of Bell's when the boy was five years of age. This was in 1872. In 1873, Sanders arranged that George should live with his grandmother, who had a big house in Salem, Massachusetts, and that Bell should board there as the most convenient arrangement to continue George's instruction.

Resumes Electrical Experiments

The arrangement included permission for Bell to use the Sanders basement for his electrical experiments, which he now resumed. In addition, he commuted 16 miles to Boston every day, taught at the University, and instructed George Sanders. Although one can well wonder where he found the time, the basement in Salem became cluttered with wires, batteries, and tuning forks. Soon, it was overflowing with apparatus and Mrs. Sanders added the entire third floor to Bell's domain.

Even before going to Boston, Bell had conceived the idea of a system of harmonic multiple telegraphy, in which a number of telegraphic signals could be sent simultaneously over the same circuit in either or both directions. This is the plan he had had in mind when he discussed acoustics with Gardiner Hubbard.

Bell's first experiments were with groups of transmitting and receiving tuning forks so arranged that each fork could vibrate between the poles of an electromagnet. The transmitting forks had connecting wires dipping into small cups of mercury so that current from a battery caused a fork to vibrate in much the same manner as an electric bell. However, each tuning fork vibrated only at its normal pitch and emitted its own characteristic musical tone. A telegraph key was placed in the circuit of each transmitting tuning fork and the group was connected through wires to the receiving forks, placed some distance from the transmitting point.

The transmitting and receiving tuning forks were arranged in pairs, that is, for each transmitter fork there was a receiver fork that vibrated at the same pitch. Each transmitter fork produced a different frequency; and, through the principle of sympathetic vibration or resonance, when the key of one of the transmitting forks was pressed, setting the fork in vibration, the fork tuned to the same frequency at the receiving end also vibrated, the other forks remaining at rest. If two transmitting forks were vibrated, the two corresponding receiving forks responded, and so on. As the forks were all connected over the same pair of wires, such a system would have made it possible to effect a considerable saving in telegraph lines.

While Bell's theory of the harmonic telegraph was sound enough in principle, he never actually succeeded in perfecting it. Practical harmonic telegraph systems were developed by others, notably Elisha Gray. Today somewhat the same principle is employed in carrier telephone and telegraph systems, voice-frequency-operated devices; and, of course, radio transmission on various frequencies to avoid interference with one another is a precise and extended application of resonance or sympathetic vibration.

There could have been many reasons why, hard as he tried, Bell did not succeed with his harmonic telegraph. First, his knowledge of electricity and mechanics was not too great. Then, he had never been particularly adept with the use of his hands. Further, his great interest in acoustics and speech naturally made the problem of "telegraphing speech" directly more intriguing than an indirect system of communication requiring code characters.

Designs Simplified Apparatus

Nevertheless, in 1873 and 1874, Bell persisted in his attempts to build a practical harmonic telegraph. From tuning forks he turned to tuned vibrating reeds equipped with double electromagnets and, for transmitting, adjustable make-and-break contacts. The receiving vibrator, of course, required no contacts. Actually, these devices were similar to modern high-frequency buzzers. While they proved somewhat better than tuning-fork vibrators, Bell made an even simpler version using only one magnet for each vibrator. Also, the reed and the spacing between reed and magnet were made adjustable.

Develops Method of Tuning Reeds—Important to Telephone Invention

While experimenting with these instruments, Bell discovered that when the reed vibrated freely it responded to a single pitch. If the transmitted pitch differed only slightly from the pitch to which the receiving reed was adjusted, no signal could be heard. He found, however, that the reeds could be forced to respond to a band of frequencies by damping the free end of the reed. This he accomplished by pressing the reed firmly against his ear while the transmitting instrument was in operation. Then, plucking the reed with his finger, he noted the pitch of the sound produced by its free vibration. If the sound, when the vibrations of the reed were damped, was higher or lower in tone than the normal pitch of the free reed, he raised or lowered the free pitch by shortening or lengthening the vibrating portion of the reed; his object in either case being to bring the free period of the reed into unison with the vibrations of the transmitting instrument. This practice was to have an important bearing on the invention of the telephone.

In all of his experiments thus far, Bell had connected his transmitters and receivers directly into the line, using two wires between the transmitting and receiving stations to form a complete circuit. However, telegraph apparatus operated with a single wire and a ground return. The resistance of such a circuit was too high to permit operation of the simple battery circuits Bell was using and, if two wires were required in the harmonic telegraph system, the advantage of such a system would be greatly reduced. About December, 1873, the idea occurred to Bell of operating his harmonic telegraph by causing the transmitting interrupters each to make and break the primary circuit of an induction coil, the secondary circuit of which could be placed in the main line by depressing a key. This arrangement permitted use of a ground return, but the induction coils introduced a new difficulty. Whereas in the first instance Bell was sending a simple, interrupted direct current between transmitter and receiver, now he was producing a ragged alternating current that the receiver reed could not readily follow.

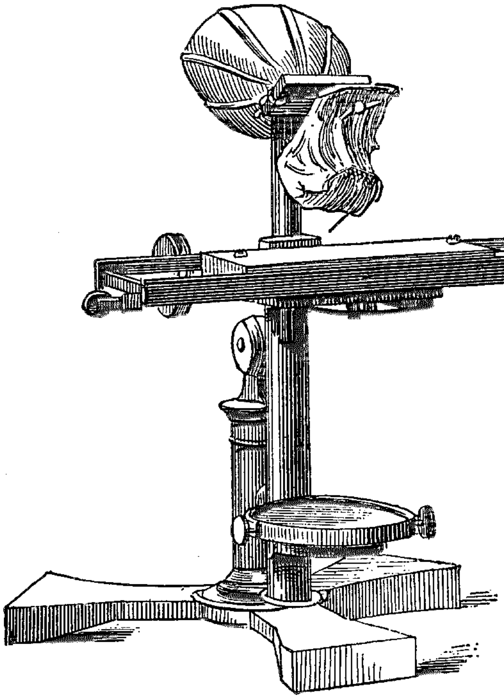
Bell considered and tried various methods of overcoming this difficulty. It occurred to him that a permanently magnetized reed, if caused to vibrate by mechanical means, could itself occasion electrical impulses of the kind required in the coils of its electromagnet. He planned to set his transmitting reeds in vibration by directing a current of air on them as in the case of organ reeds. But he doubted whether an electric current generated by the vibration of a magnetized reed in front of an electromagnet would be powerful enough to produce at the receiving end of the circuit a vibration sufficiently intense to be practical for use on real lines for multiple-telegraph purposes. So, instead of pursuing the idea further at that time, he devised a rotary circuit breaker to interrupt the current rapidly enough to provide intermittent currents equal in frequency to that of the receiving reed.

The Phonautograph and Manometric Capsule

During the winter of 1873–1874, while Bell was experimenting with his harmonic telegraph, he also investigated the possibilities of two devices then in use in the physics laboratory at the Massachusetts Institute of Technology. They were the phonautograph and the manometric capsule. Both instruments produced visible patterns from sounds. Bell saw the possibility of utilizing them in his work of teaching speech to deaf children.

The phonautograph consisted of a speaking trumpet closed at one end by a stretched membrane to which was attached a light wooden lever bearing a bristle stylus at its far end. The point of the bristle just touched a plate of smoked glass which could be moved at a uniform rate across it. When the bristle was vibrated by the voice speaking into the trumpet, it traced a shape on the plate that varied with the sound uttered.

The manometric capsule consisted of a cavity in a piece of wood, divided in half by a diaphragm of gold-beater's skin. One side had a gas inlet and a burner connected to it, while the other compartment was connected to a speaking tube. When sound entered the tube, the air vibrations were communicated through the membrane to the gas and thence to the flame of the lighted burner. The flame, accordingly, followed the sound variations. On viewing the reflection of the flame in a mirror revolving at a suitable speed, the resulting stroboscopic effect showed definite sound patterns. For example, the vowel ē presented the appearance of a long band of light with teeth like a saw. When the sound was changed to ĭ, as in "it," each tooth of the saw became notched. The vowel ĕ, as in "bed," caused the appearance of a pleasing and complicated pattern that resembled lace.



Drawing of the human-ear phonautograph built by Alexander Graham Bell in 1874. The stylus wrote sound patterns on a plate of smoked glass which was moved across the upper platform. The above drawing appeared in the published pamphlet of Bell's lecture on the telephone presented before the Society of Telegraph Engineers, Westminster, England, 1877.

Bell hoped to use one device or the other to photograph wave patterns for his deaf pupils. He felt that, if he could produce a clearly defined pattern of sound, for instance the sound of ē, then the deaf child could practice uttering the sound into a phonautograph or manometric capsule until he could duplicate the pattern with certainty. However, neither device produced useful patterns.

Human-Ear Phonautograph

Bell noticed that the phonautograph somewhat resembled the structure of the human ear. Once again returning to Nature, as he did when he built the speaking machine, Bell decided to study the human ear mechanism more closely with the idea of constructing a phonautograph more like an ear. He, therefore, discussed the problem with Dr. Clarence J. Blake, a noted Boston ear specialist. Blake suggested that Bell attempt to construct a mechanism from an ear taken from a dead person. Dr. Blake prepared a specimen for Bell who used it successfully in making tracings of sound vibrations on smoked glass.

The human-ear phonautograph never produced patterns suitable for teaching speech to deaf children, but it had two qualities that were directly instrumental in leading Bell's thinking toward the invention of the telephone. First, Bell marvelled at the ability of the tiny ear-drum diaphragm to move the proportionally huge bones of the inner ear, as a result of the vibration of sound waves. The evidence of this motion was translated by the phonautograph into a visible pattern right before his eyes. Further, since the phonautograph was a crude form of oscillograph, he could see that sound waves followed a pattern that was undulatory in character. It has been said that, in relating this point to his knowledge of acoustics, and then further relating it to the undulating pattern electrical currents would have to follow to reproduce sound, Bell hit upon the basic factors that made the electric telephone possible. In the light of present knowledge, where these relationships are commonplace, the importance of Bell's reasoning and discovery might easily be missed. But considered in the light of the historical facts, it has been proved that none of Bell's contemporaries, attempting to produce a talking telegraph, employed the principle of the undulatory current.

Further Concepts of Undulatory Current

Bell's experiments with the human-ear phonautograph were performed during the summer of 1874 while he was at Brantford, Ontario, visiting his father during a vacation period. In the course of that summer, in connection with his work on the harmonic telegraph, his thoughts went back to the idea of mechanically vibrating a polarized reed in front of a magnet. He came to realize that the vibration of the reed would produce on the line wire an undulatory current of electricity corresponding to the undulatory motion of the reed. That is, the direction of polarity of the induced current sent over the line would correspond to the direction of the motion of the reed, being positive when the reed moved in one direction and negative when it moved in the other. Also, the intensity of the current would correspond, from instant to instant, to the velocity of the movement of the reed, being greatest when the reed was moving fastest.

Bell next considered the effect of two or more reeds set into vibration simultaneously. He could see that, when two reeds were in phase, their currents would aid each other, while when they were out of phase the currents might be opposed. He then concluded that [4]

> the resultant electrical effect produced upon the line wire by the simultaneous vibration of both reeds would then be expressed by a curve representing the algebraical sum of the two curves considered.

Studying this effect further, Bell saw that a similar result would follow if he had a multitude of reeds of different pitch, each with its own coil, and all the coils connected in one circuit. The thought then occurred to him of having a single electromagnet for all reeds instead of a separate coil for each reed. This line of thought led Bell to the conception of the "harp" apparatus.

In the harp apparatus, identical instruments were to be used at each end of the line and each instrument was to be capable either of transmitting or receiving. Two series of steel reeds, each having a different pitch, and all polarized from a single permanent magnet, were to be placed opposite the pole pieces of a wide electro-

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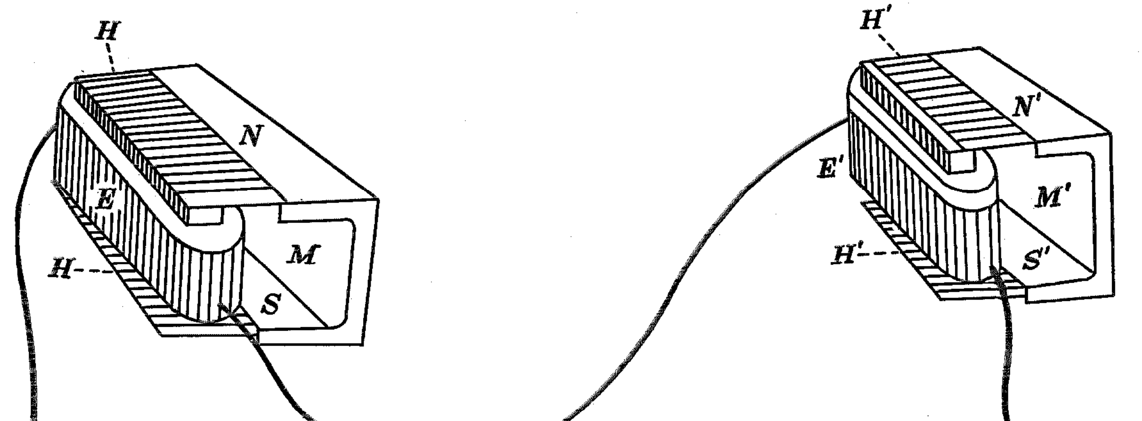
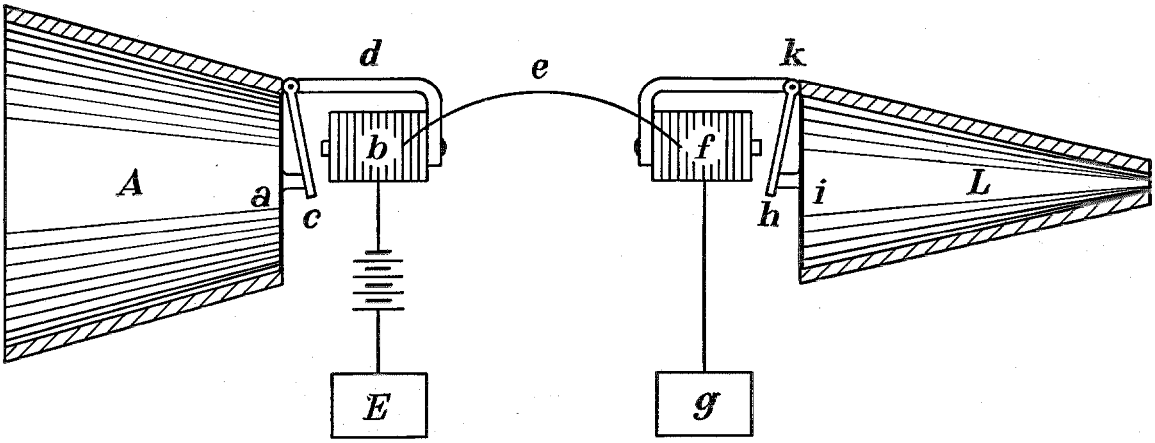
[4] Deposition, Circuit Court of the United States, District of Massachusetts. No. 2356A, page 33. Printed by the American Bell Telephone Co., 1908, as the "Red Book."



Bell's conception of his "harp" telegraph apparatus. *H, H* were two series of steel reeds, each reed having a different pitch. A single permanent magnet *M* was to be employed. An elongated electromagnet *E* was to be placed between the two series of reeds so that the free end of each reed might be close to the pole of the electromagnet without touching it. Bell never reduced this idea to practice, because it became evident to him that a single diaphragm could function as well as the many reeds.

magnet so that the free end of each reed would be close to one of the poles of the electromagnet.

The concept of the harp apparatus brought him back to the phenomenon which he had originally used to illustrate the idea of the harmonic telegraph, the sympathetic vibration of piano strings when notes are sung close to them. In like manner, if a harp had enough reeds, it should pick up every sound of the voice. One could talk into it, and the vibrations would be carried by a complex current to a similar receiver which then would vibrate in such a way that the sound would be repeated. Bell never built the harp apparatus. He felt it would be too complicated to be practical, and, at the time, it would have been too expensive for his limited resources.



Copy of the drawing of "Figure 7" from Bell's patent of the telephone issued March 7, 1876. The transmitter is shown at the left and the receiver at the right. In actual practice, Bell found that equipment similar to this would transmit speech without the battery shown connected to the transmitter.

The Membrane Speaking Telephone

Since Bell was not actually attempting to build the harp apparatus that summer in Canada, he was left free to speculate on further refinements. He knew that the manometric capsule and the phonautograph, employing a single diaphragm, were able to reproduce graphically resultant sound patterns of all vibrations impressed on them, regardless of their complexity. He concluded that a single reed, if it could be made to vibrate like these diaphragms, would generate the same complicated currents as an infinite number of reeds on the harp.

From this line of reasoning came the conception of the membrane speaking telephone. Before he left Brantford, Bell had worked out the plan for such a telephone. His plan was to attach the free end of a reed to the center of a stretched membrane, but instead of firmly fastening the reed to the pole of a permanent magnet, the magnet pole was simply to be brought into close proximity to the hinged end of the reed, so as to polarize it by magnetic induction. The arrangement thus conceived was substantially similar to that subsequently shown as "Figure 7" of his U. S. patent 174465 of March 7, 1876.

While Bell was satisfied that the apparatus he had conceived would produce voice currents, he doubted that strong enough electrical currents would be generated to be heard at the end of a real line. Since experimentation was costly in time and funds, he decided to return to the apparently more practical task of the harmonic telegraph. He did not, therefore, construct the projected telephone when he returned to Boston.

Although Bell received some income from teaching, it is doubtful if he would have been able to proceed so rapidly with his work if he had not obtained financial help. This aid he began receiving in the fall of 1874 from both Hubbard and Sanders. The two men had separately volunteered to give Bell financial assistance. So Bell, somewhat embarrassed by two such offers, brought them together. At this meeting Hubbard and Sanders agreed to share equally in Bell's expenses, which he put at an extremely modest figure. They also agreed at the time that the shares in any resulting patents would be divided equally three ways.

Bell Meets Thomas A. Watson

Bell began by building his own equipment, but, at length, it became clear to him that he lacked both the skill and the time to do this work properly. So he took some of his apparatus to the shop of Charles Williams, Jr., at 109 Court Street, Boston, to be refashioned by an expert electrician. The young man assigned to this work was Thomas A. Watson. Later, Bell asked Watson to leave the Williams shop and come to work for him. They not only became fast friends, but after a time Watson was made the fourth member of the group to share in the patents as part payment for his work.

By February 1875, when Bell was only 28 years old, his harmonic telegraph had advanced to the point where he thought patent applications should be filed on it. Consequently, during that month, he put off all pupils and classes and visited Washington, D. C., to draft specifications and file three patent claims. While there he saw Joseph Henry, then secretary of the Smithsonian Institution. Henry saw Bell's hastily set up telegraph apparatus and Bell talked to the aged scientist about his telephone theory and asked if he should publish it. Henry advised against it, telling Bell that he had the germ of a valuable idea. Bell replied that he feared he lacked the electrical knowledge to work it out. Henry simply said, "Get it!" and encouraged him in other ways. Later Bell wrote, "But for Joseph Henry, I should never have gone on with the telephone."

Bell received a patent on his harmonic-telegraph apparatus a few days after his talk with Henry. While still in Washington, Bell demonstrated the apparatus to William Orton, president of the Western Union Telegraph Company. Western Union was then the largest corporate body in the U. S. A. Consequently, it was the logical and, perhaps, the only customer in this country for any telegraph invention. Moreover, the telegraph company's wires were known to be terribly overloaded with messages and it was generally understood that Western Union would pay as much as a million dollars for a device capable of increasing the number of messages that could be sent over a single line in a given time.

It was natural that a practical business man like Hubbard, under these circumstances, would prefer to see Bell develop the harmonic telegraph rather than the telephone, which was regarded even years after it became a reality as a toy or, at most, a diverting lecture subject. So, despite all of Bell's dreaming about the telephone, the insistence by Hubbard and others that he continue with the harmonic telegraph could not be ignored. Besides Bell, who was greatly in need of funds, saw the possibility of selling the telegraph patents for a large sum which would provide for his experimenting to his heart's content with the telephone.

Bell was heartened when the demonstration before Orton in Washington went well and he received an invitation to bring the harmonic telegraph to Western Union's laboratory in New York City. He was told by Western Union engineers in New York that the apparatus still had bugs and he received a number of suggestions for making improvements.

Transmission of Sound

When Bell returned to Boston, he had Watson make up three sets of multiple-telegraph apparatus of the make-and-break variety employing tunable reeds. These were set up in the attic of the Williams building, where Bell had obtained the use of two adjoining rooms; a wire line was strung between them.

On June 2, 1875, Bell and Watson were ready for a test of this apparatus. Bell, at one end of the line, was tuning up the metal reeds on a group of receivers. Watson, in the adjoining room, was sending the tones of the transmitters to Bell. What followed was probably the most crucial moment in the invention of the telephone. That moment was described by Watson as follows:[5]

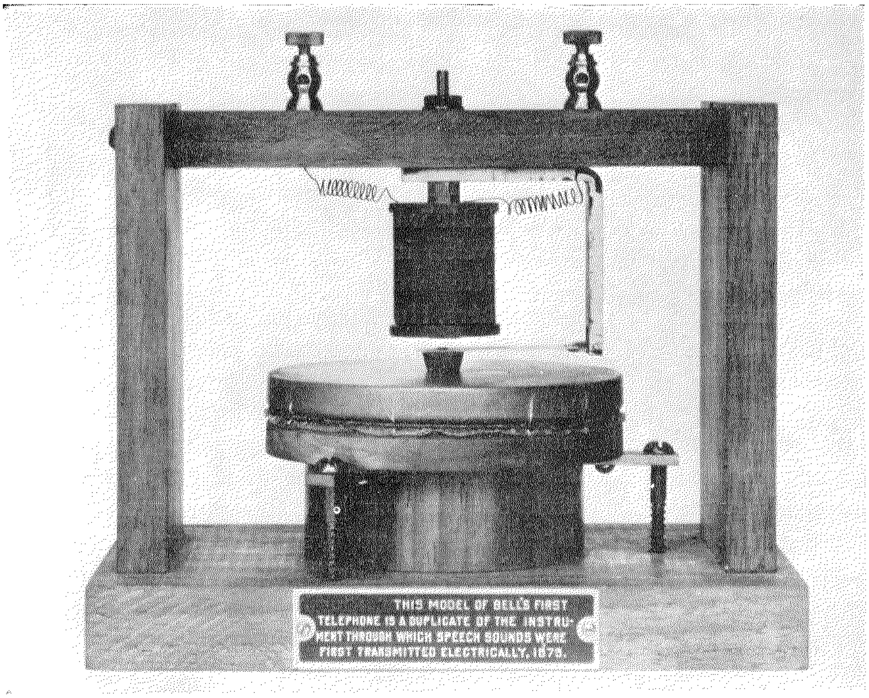
> I had charge of the transmitters as usual, setting them squealing one after the other, while Bell was retuning the receiver springs one by one, pressing them against his ear. . . . One of the transmitter springs I was attending to stopped vibrating and I plucked it to start it again. It didn't start and I kept on plucking it, when suddenly I heard a shout from Bell in the next room, and then out he came with a rush, demanding: "What did you do then? Don't change anything! Let me see?" I showed him. It was very simple.

The make-and-break points of the transmitter spring Watson was trying to start had accidentally been brought into permanent contact, so that when he snapped the spring the circuit had remained closed. Of this, Watson said:[5]

> . . . that strip of magnetized steel, by its vibration over the pole of its magnet, was generating that marvelous conception of Bell's—a current of electricity that varied in intensity precisely as the air was varying in density within hearing distance of that spring. That undulatory current had passed through the connecting wire to the distant receiver which, fortunately, was a mechanism that could transform that current back into an extremely faint echo of the sound of the vibrating spring that had generated it, but what was still more fortunate, the right man had that mechanism at his ear during that fleeting moment, and instantly recognized the transcendent importance of that faint sound thus electrically transmitted. The shout I heard and his excited rush into my room were the result of that recognition. The speaking telephone was born at that moment.

Bell, in holding the harmonic receiver tightly against his ear, effectively clamped the free end of the reed or spring, thus damping its natural rate of vibration and causing it, instead, to vibrate in a manner analogous to the diaphragm of the modern telephone receiver. Instead of the customary whine of the intermittent battery current, he had heard an unexpected sound—the twang of a plucked reed—a tone with overtones.

What removes this occurrence from the realm of chance discovery is the fact that Bell, who not only had reasoned that such a current could be generated but foresaw the results it would produce, was probably the one man in the world at that time qualified by training and acuteness of ear to recognize in the faint sound that came to him a confirmation of his theories.


THIS MODEL OF BELL'S FIRST
TELEPHONE IS A DUPLICATE OF THE INSTRUMENT THROUGH WHICH SPEECH SOUNDS WERE
FIRST TRANSMITTED ELECTRICALLY, 1875.

It was an instrument like this that Thomas A. Watson made for Alexander Graham Bell overnight from June 2 to June 3, 1875. It transmitted speech sounds, thus encouraging Bell to go on with development of the telephone.

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[5] Thomas A. Watson, "The Birth and Babyhood of the Telephone," *Proceedings of the Telephone Pioneers of America*, p. 36; Oct. 17, 1913.

As Bell subsequently wrote about the event in the language of a patent suit deposition:[6]

> These experiments at once removed the doubt that had been in my mind since the summer of 1874, that magneto-electric currents generated by the vibration of an armature in front of an electromagnet would be too feeble to produce audible effects that could be practically utilized for the purpose of multiple telegraphy and of speech transmission.

First Voice Sounds Received

Before that history-making day was ended, Bell and Watson had tried and retried the same experiment many times. The next step was to construct the first speaking telephone. Bell gave Watson directions, which followed closely the design of the membrane telephone he had conceived in Brantford the summer before. One of the harmonic receivers was to be mounted in a wooden frame, the free end of its spring to be fastened to a small bit of cork attached to the middle point of a drumhead or diaphragm of tightly stretched parchment also mounted in the wooden frame. There was to be a mouthpiece for concentrating the voice waves on the opposite side of the diaphragm.

The next day, June 3, 1875, Watson built the first electric speaking telephone. It had many deficiencies, but before that day was ended enough faults had been corrected so that it could transmit the sound of Bell's voice to Watson. Because the energy generated by the transmitter was weak and the receiver was insensitive, Watson could not hear words, just recognizable voice sounds. Today, however, an exact replica of this first telephone, when connected as the microphone of a public address system, transmits perfectly clear and intelligible speech with a slightly boomy resonance. Connected as a receiver to the output of an amplifier, reproduction is clear enough, but weak and predominant in lower-register frequencies.

Bell and Watson went on experimenting with the telephone all summer. In September, while again at Brantford, Bell started writing the specifications for his first telephone patent. The claims thus written ultimately resulted in his basic U. S. patent 174465 granted on March 7, 1876. From the draft which he wrote and rewrote, resulted the important fifth claim of that patent:

> The method of, and apparatus for, transmitting vocal or other sound telegraphically, as herein described, by causing electrical undulations, similar in form to the vibrations of the air accompanying the said vocal or other sounds, substantially as set forth.

Bell had spent about five months casting and recasting the patent specifications. During these months he had sought a method of putting a stronger undulating current on the line than was possible with magnetic induction. He had decided this could be done by causing a resistance to fluctuate, so that a current through it would be stronger and weaker as the transmitter diaphragm vibrated. After considering several means of providing such a variable resistance, he decided to use a wire, working like a plunger, up and down in a conducting fluid. So Bell included in his patent specification the description of a transmitter wherein—as the voice made the transmitter vibrate—the wire attached thereto would move up and down in a fluid conductor. As it went deeper, it lessened resistance; as it rose again, the resistance increased. Current passing through the wire and fluid would thus undulate as the sound waves varied.

Financial Difficulties

The winter of 1875–1876 was a difficult time for Bell. He wanted to get on with his telephone and develop it to the point where it actually transmitted words. But he felt an obligation to his backers, Hubbard and Sanders, and Hubbard still favored the harmonic telegraph. Unfortunately, in the agreement they had made, Bell had neglected to request funds for himself and now needed money. Consequently, he gave a series of lectures in the fall of 1875. About this time he also completed arrangements with the Honorable George Brown of Toronto, Ontario, and his brother Gordon to pay the expenses of private rooms for housing his electrical apparatus. He felt he needed these rooms because he had heard of strangers visiting the Williams shop and examining his apparatus. With funds provided by the Browns, he engaged two rooms at a boarding house at 5 Exeter Place, Boston,

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[6] Deposition, Circuit Court of the United States, District of Massachusetts. No. 2356A, page 59. Printed by the American Bell Telephone Co., 1908, as the "Red Book."

Massachusetts, to which he moved his apparatus in January, 1876.

He then gave up his lectures and devoted his entire time to experimenting. Watson continued to construct his apparatus and assisted in testing it. Bell's zeal at this time has been attributed to his desire to marry Mabel Hubbard. He knew that his professional work, if continued, could yield him a modest income; but he was confident that, if he devoted his attention exclusively to his electrical inventions, they would bring him a fortune—and a fortune was the only thing the modest Bell cared to offer Mabel, daughter of the wealthy Gardiner Greene Hubbard.

First Sentence Received

The first experiments at the Exeter Place rooms were, strangely enough, not with the telephone, but with the harmonic telegraph which Bell was still trying to perfect and sell to Western Union. At the time he was working on an electric spark arrester for the contacts of the make-and-break circuits of the transmitters. This was so similar to the variable-resistance transmitter he had described in his patent specification that out of it he developed such a transmitter.

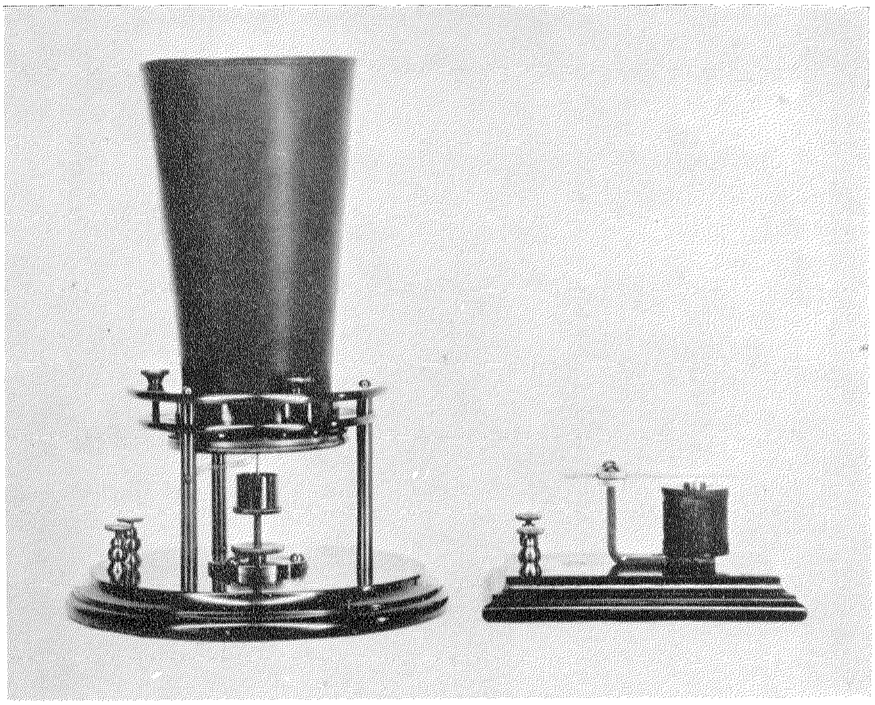
It was on March 10, 1876, three days after the telephone patent was granted to Bell, that the first complete and intelligible sentence ever transmitted by telephone was heard. The circumstances were as dramatic as any connected with Bell's previous efforts to transmit speech. Watson had built a variable-resistance transmitter and both Bell and he were planning to spend the night at Exeter Place testing it. Neither of them had any idea that they were about to use the best transmitter yet devised. They diluted the sulphuric acid for its cup, connected it to the battery and to the wire running between the rooms, and then Watson went to the other room to listen.

Almost immediately Watson heard Bell's voice come shouting from the receiving instrument, "Mr. Watson, come here, I want you."

It was a call for assistance. Bell had spilled some battery acid over his clothes as he completed setting up the test transmitter. Since it was only a one-way line, Watson ran to Bell's end of it to shout, "Mr. Bell, I heard every word you said—distinctly." The electric telephone had spoken at last, and so clearly that doubt no longer existed as to its capability of transmitting articulate speech.

Iron-Box Receiver

Bell and Watson continued to improve their apparatus and it was about that time that Bell designed the receiver that was to be the forerunner of all telephone receivers. He called it the



Replicas of the variable-resistance transmitter and reed receiver. It was over equipment such as this that Bell shouted, "Mr. Watson, come here, I want you." This was the first intelligible sentence received over the telephone.

"iron-box" receiver. It employed a metal diaphragm and operated on exactly the same principle as the telephone receiver in use today.

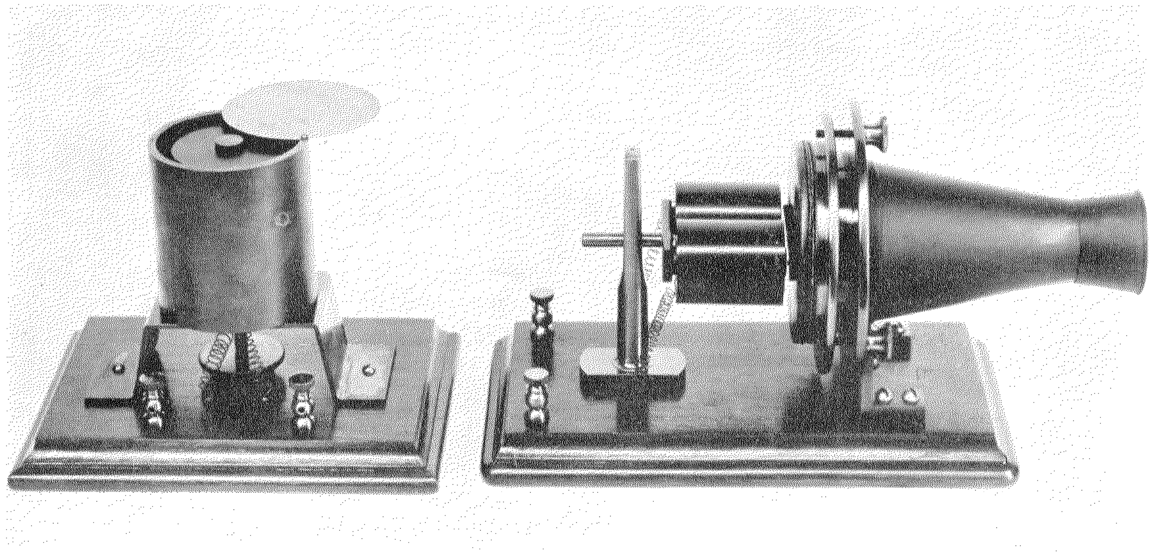
As already described, it had been Bell's habit, while using a tuned-reed receiver, to prevent free vibration of the reed by pressing it closely against his ear. It occurred to him to overcome this difficulty by fettering the free end. At first he proposed clamping both ends of the reed and attaching them to the outer two poles of a W-shaped electromagnet. He soon concluded, though, that a cylindrical iron box with a central core would be better. As he worked it out, the end of the central core became one pole of the tubular magnet thus formed and the rim of the iron box the other pole. By placing a lid or diaphragm on the end of this magnet, he had an armature that was damped all around its periphery and polarized by contact with the rim of the box. The iron-box receiver worked better than any he had used previously. It was this receiver that he demonstrated at the Philadelphia Centennial Exposition in June, 1876.

Tries to Sell Patents for $100,000

Bell now had apparatus that worked effectively. He had a patent covering the telephone. But he had a long struggle ahead of him. First, he had to prove that the telephone would operate over long stretches of wire. Then, he had to convince the skeptics that the telephone was more than a scientific toy. People generally refused to believe that transmitting the voice over wires was possible. When they actually heard the telephone work, they thought it was accomplished by trickery. At one time, Bell was so discouraged that he offered his patents to Western Union for $100,000. Fortunately for him, Western Union turned him down. It has been said that a few years later Western Union would have been glad to pay twenty-five million dollars for the Bell patents.

Bell's magnificent speaking voice and his experience as a lecturer served him well during these discouraging years. He gave numerous lectures and demonstrations and actually "sold" the telephone to the public.

These efforts kept Bell from new experiments for several years. Shortly after his return from his honeymoon in England, he began to relinquish to others the task of improving the telephone and developing the necessary associated apparatus. Throughout his life, however, he maintained his interest in scientific matters covering a wide range of subjects.



Replicas of the "iron-box" receiver and an improved magnetic transmitter. The iron-box receiver was developed by Bell shortly after the first intelligible speech was transmitted over the telephone. It employed a metal diaphragm and permanent magnet in a form comparable to the most modern receivers used in telephones today.

Founds Volta Laboratory

In 1880, he perfected and patented the photophone, a device for transmitting speech over a beam of light. Bell used the well-known principle that the resistance of crystalline selenium varies with change in light intensity. Little or no practical return came to Bell from his invention. It is interesting to note, however, that this same principle of transmitting speech over light rays was used by both sides for communication purposes during World War II. Also, at present, experiments are under way with a view to using infra-red rays, as Bell used them in his photophone, for applications such as the relaying of television programs.

An offshoot of his work with the photophone, Bell called the spectrophone, which he described in a paper before the National Academy of Sciences on April 21, 1881. He had noted in the course of his experiments that many substances, when subjected to an intermittent beam of light, would give off a tone in a telephone receiver. He drew attention to the use of such equipment, particularly, in the detection of invisible rays. As far as is known, nothing came of these experiments.

In 1880, Bell received the Volta prize of 50 000 francs from the French government for his invention of the telephone. With the money, he established the Volta Laboratory at Washington, D. C., in collaboration with Sumner Tainter, a maker of optical instruments, and Chichester A. Bell, a cousin who was a specialist in organic chemistry. Each of the three was to work on his own line of pure research. However, when funds began to run out, they decided to take up some line of applied science jointly to finance the institution.

Volta Group Develops First Disk Phonograph Records

They developed an improvement in the making of records for Thomas A. Edison's phonograph which, invented in 1877, was still using metal foil as a recording medium. The Volta group took out the basic patents for phonograph recording on wax cylinders and disks, and sold these to an operating company that launched the recording industry as it is known today. The original idea for using a disk was to permit voice recordings to be mailed conveniently. They were intended as a substitute for written messages, a practice which has only begun to come into use recently.

Bell put his share of the proceeds of the sale of the recording patents into a branch of the laboratory he named the Volta Bureau, set up specifically to carry on his work for the deaf. The Bureau at that time was making a statistical study of deafness, particularly of the probability of inherited deafness. The Volta Bureau, still at Washington, continues its work to this day in behalf of the deaf.

Electric Probe

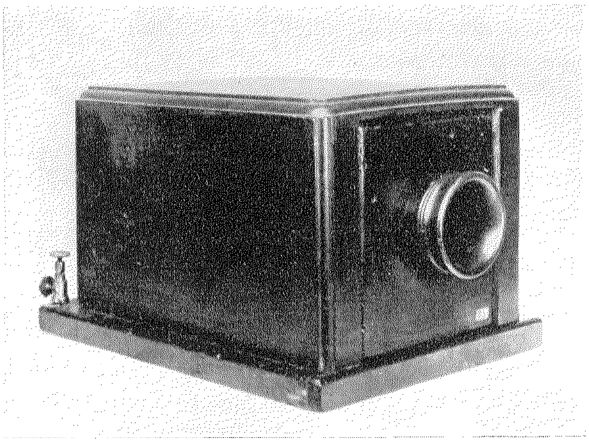
When President Garfield was shot in 1881, Bell, who was then living in Washington, hit upon the idea of locating the bullet with an induction balance. In the heat of that Washington summer, Bell labored day and night to test the ability of the induction balance to locate metallic masses in the human body. However, Bell's attempts to locate the bullet in the president failed because the doctors in attendance, although trying to comply with Bell's request to remove all metal from Garfield's surroundings, overlooked a steel bedspring directly beneath him. When discovered, further tests were hampered by the fact that Garfield's condition made it difficult to move him. After Garfield's death, Bell perfected an electric probe which was used in surgery for several years and for which he received an honorary degree of Doctor of Medicine from Heidelburg University in 1886. X-ray technique supplanted the probe about the turn of the century.

Stops Work on Telephone

Although Bell continued scientific research for the rest of his life, it was about this time that he stopped further work on the telephone and other electroacoustic devices. One reason may have been the discouraging round of patent suits he and his associates had to face. Although Bell's right to the telephone patent was upheld by the highest courts, his sensitive nature recoiled from such experiences.

Another explanation is given in a letter written by Mrs. Bell shortly after his death:

> I verily believe (she wrote) that the reason Dr. Bell did not follow up his invention of the photophone—the reason he did not follow that up, and the reason he took up aviation instead was that I could not hear what went on over the radiophone (as the photophone was called later) but that I could see the flying machine.



The first commercial telephone was this wooden-box model, used in 1877. It was soon supplanted by other forms of boxes, then by hand telephones turned out of wood, that resembled closely the receivers later used on desk telephones.

General John J. Carty, who received this letter, commented as follows:[7]

> It was a very touching letter and reminds me of what happened on his great day of triumph when Bell talked across the continent to Washington and when he received the congratulations of the Chief Magistrate of our Nation. Mrs. Bell was there and Bell, with poignant sadness, looking toward his wife, said to me, "And to think that she has never heard through the telephone."

Bell had great faith in the future of aviation and spent much of his time in research on this subject. All his life Bell had been interested in flight, and his support to aviation was given at a time when to do so risked his scientific reputation. In 1891, he contributed $5000 for Langley's aviation experiments and on May 6, 1896, saw the successful flight of Langley's steam-driven 16-foot model which, however, did not carry a man.

Elected Regent of Smithsonian Institution

When, in 1898, Bell was elected a Regent of the Smithsonian Institution, his enthusiasm for Langley's experiments had much to do with obtaining from the War Department an appropriation of $50,000 to be used by Langley for the development of aeronautics.

In 1898, he became president of the National Geographic Society and served in that capacity until 1903. He helped finance the society and was instrumental in building its magazine into a national institution.

Much of Bell's later life was spent at his estate near Baddeck, on Cape Breton Island, Nova Scotia. He acquired the estate after a visit there in 1885 and named it Beinn Bhreagh. It was at Beinn Bhreagh that he set up a laboratory and carried on many experiments. While there, he kept a staff of experimenters busy on such diverse ideas as devising a means of condensing fog to furnish fresh water for men adrift at sea, and attempting to breed sheep that would bear more than one lamb at a time.

Varied Scientific Interests

Besides numerous speeches and papers on subjects related to training the deaf to speak and the telephone, Bell wrote articles on the photophone, spectrophone, medical and surgical subjects, eugenics and longevity (based on his studies of the census), his experiments in sheep breeding, aerial locomotion, as well as many general subjects.

The breadth of his interests and the originality of his thinking are indicated by the fact that in 1882 he published "A Proposed Method of Producing Artificial Respiration by Means of a Vacuum Jacket" in which he describes a device anticipating today's iron lung. In 1885, following the collision of a steamer with an iceberg, he wrote advocating a method of determining the location of such an object at sea by detecting a sound echo from it. In a talk in 1906, he advocated measuring the depth of the ocean by echo. Toward the end of his life, he interested himself in the heating and ventilating of houses and, when he had to spend some weeks in Washington, D. C., one hot summer, he air-conditioned his study by directing an electric fan over cakes of ice.

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[7] John J. Carty, "Episodes in Early Telephone History," *Bell Telephone Quarterly*, v. 5, pp. 66–67; April, 1926.

Honors

Among the honorary degrees conferred on Alexander Graham Bell were: Doctor of Laws: Illinois College, 1881; Harvard College, 1896; Amherst College, 1901; St. Andrew's University, 1902; Edinburgh University, 1906; Queen's University, Canada, 1908; George Washington University 1913; Dartmouth College, 1914. Doctor of Philosophy: National Deaf-Mute College (now Gallaudet College), 1880; Wurzburg University, 1882. Doctor of Science: Oxford University, 1906. Doctor of Medicine: Heidelberg University, 1886.

In addition to receiving the Volta prize in 1880, he was made an officer of the Legion of Honor by the government of France in 1881.

The medals awarded Bell were: Centennial Exposition, Philadelphia, 1876, gold medal for speaking telephone and gold medal for visible speech. Royal Cornwall Polytechnic Society, 1877, James Watt silver medal for the telephone. Massachusetts Charitable Mechanics Association, 1878, gold medal for the telephone and gold medal for visible speech. Society of Arts, London, 1878, Royal Albert silver medal for his paper on the telephone. Rêpublique Française Exposition Universelle International, Paris, 1878, gold medal for the telephone and a silver medal. Society of Arts, London, 1881, Royal Albert silver medal for his paper on the photophone. The Karl Koenig von Wuertemberg gold medal. Society of Arts, London, 1902, Royal Albert gold medal for his invention of the telephone. John Fritz gold medal, 1907. Franklin Institute of Philadelphia, 1912, Elliott Cresson gold medal for the electrical transmission of speech. David Edward Hughes gold medal and a silver medal, 1913. American Institute of Electrical Engineers, 1914, Thomas Alva Edison gold medal.

Alexander Graham Bell's life ended on August 2, 1922, at Baddeck, Nova Scotia, where his body was laid at rest on the summit of a near-by hill.

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J. E. Kingsbury, "The Telephone and Telephone Exchanges—Their Invention and Development," Longmans, Green, and Co., London and New York, 1915.

Catherine Mackenzie, "Alexander Graham Bell—The Man Who Conquered Space," Houghton Mifflin Co., Boston and New York, 1928.

Frederick L. Rhodes, "Beginnings of Telephony," Harper and Brothers, New York and London, 1929.

"The Career of Alexander Graham Bell," American Telephone and Telegraph Company, New York, 1946.

France-England Submarine Cable (1939) and Paris-Calais Cable*

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THE France-England submarine cable (1939) and the Paris-Calais cable are the first cables of French manufacture to include 12-channel telephone circuits; it also appears that they are the first cables in which the two transmission directions of a 12-channel carrier link were provided under the same lead sheath.

In this paper, the problems presented by the transmission of 12-channel carrier telephony over cable circuits is first reviewed; particularly the requirements for the Paris-Calais cable, as set forth in the special specification for it. The cable structure is then given in greater detail and some aspects of the manufacture of the submarine cable are described. Finally, the results obtained and the methods used are discussed.

The calculation of near-end cross-talk compensation between circuits having different propagation constants is developed in an appendix.

* * *

Communication between Paris and northern France has and continues to play a very important part in that country's economic life. In 1937, there were three cables for telephone conversations: a Paris-Boulogne cable (1927) and two Paris-Lille cables (1926 and 1933). However, the number of circuits available was still inadequate for handling the traffic requirements; consequently, the manufacture and installation of a new cable, the Paris-Calais-Saint Margaret cable, was considered. On the one hand, the circuit length and the Paris-London traffic density justified the use of 12-channel telephony, and, on the other hand, the accommodation of such important centers as Clermont, Amiens, and Calais required 2- and 4-wire circuits usually included in interurban telephone cables. The French Telephone Administration, therefore, decided to group in this cable both loaded circuits for normal 2-wire working (H.88/36) and 4-wire working (H.22/9), and nonloaded circuits for 12 telephone channels used for Paris-London through connections.

The Sangatte-Saint Margaret cable was made in 1938–1939 and laid in July, 1939; then came the Paris-Calais cable, the manufacture of which began in 1939, and the through-splicing finished in 1941. The Paris-Calais cable is at present only partially used, circumstances having prevented the installation of equipment for 12-channel operation. These are the first cables of French manufacture which include 12-channel circuits. Therefore, some detailed information will be given concerning them.

1. *Transmission Requirements for 12-Channel Circuits*

1.1 TRANSMISSION PROBLEM

The problem is to transmit the same band of frequencies (48 kilocycles per second) over a number of pairs, some of the pairs being assigned to transmission in one direction, the other pairs to the opposite direction. The choice of frequency band is the result of a compromise. To facilitate the manufacture of repeaters (amplifiers), it is most convenient to use a relatively high frequency and thus decrease the bandwidth as a percentage of the transmission frequency. However, the circuit attenuation increases with frequency and this suggests the use of low frequencies. The band from 12 to 60 kilocycles was finally chosen, the attenuation between two successive repeater points being some 6 to 7 nepers.† Table I gives a comparison of these values with the corresponding values for the most commonly used loaded circuits.

The transmission of such a wide frequency band raises a problem of far-end cross talk. The operation of circuits with such high attenuation adds a near-end cross-talk problem.

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* Originally published under the title, "Le câble sous-marin France-Angleterre 1939 et le câble Paris-Calais," *Revue Générale d'Electricité*, v. 54, pp. 271–280; September, 1945.

† 1 neper = 8.686 decibels.

Far-end cross talk had already been studied for cables laid in the United States,[1] in England,[2] and in the Netherlands,[3] but near-end cross-talk difficulties between circuits transmitting in opposite directions had been avoided by placing these circuits in separate cables. For the Paris-Calais-Saint Margaret link, however, it was decided to include in one cable, i.e., under the same lead sheath, the 12-channel circuits of opposite directions. The solution of this problem is of particular interest for submarine cables, as there is an obvious advantage in not increasing unnecessarily the number of cables submerged in the Dover Strait; it is also of interest for land cables as it reduces the cost of the cable sheath and simplifies laying.

Experience with the previous cables[1,2,3] made it possible to determine the requirements for insuring an acceptable transmission quality with 12-channel pairs. These conditions are relative, first, to the electrical characteristics of manufactured cable lengths and, second, to the transmission characteristics of the cable sections between two repeater points. Requirements relative to conductor resistance, resistance unbalance, insulation resistance, leakage, and voltage breakdown are the same as for interurban cable circuits and will not be discussed. Only those characteristics especially applicable to 12-channel transmission will be considered.

TABLE I

TRANSMISSION CHARACTERISTICS OF TELEPHONE CIRCUITS

Type of Circuit	Frequency Band Assigned to Telephone Work (Kilocycles)	Attenuation Between Repeater Points (Nepers)
2-Wire H.88/36	0.3–2.7	1–2
4-Wire H.22/9	0.3–5.7	2.5–3.5
12-Channel Nonloaded Pair	12–60	6–7

TABLE II

CAPACITANCE UNBALANCES MEASURED AT 800 CYCLES

Circuit	For 100 Percent of the Lengths		For 90 Percent of the Lengths	
	Average ($\mu\mu$f)	Maximum ($\mu\mu$f)	Average ($\mu\mu$f)	Maximum ($\mu\mu$f)
Between side circuits of the same quad	46	180	34	135
Between pairs of adjacent quads Between adjacent pairs	14	90	14	90
Between pairs of nonadjacent quads Between nonadjacent pairs	—	30	—	30
Between the two wires of one pair and ground	180	670	140	500

1.2 ELECTRICAL CHARACTERISTICS OF CABLE LENGTHS

The values given below are those of the Paris-Calais cable specification. This cable was manufactured in lengths of about 230 meters (750 feet); the values for the submarine cable are slightly different because that cable was manufactured in lengths of 400 meters (1300 feet).

Diameter of conductors = 1.3 millimeters.

A. Effective capacitance (average of pair capacitances for one direction of transmission):

$35 \cdot 10^{-9}$ farad per kilometer ±5 percent for 90 percent of the lengths manufactured.

$35 \cdot 10^{-9}$ farad per kilometer ±8 percent for 100 percent of the lengths manufactured.

B. Difference between the capacitance of one pair and the average of the capacitances of all pairs assigned to the same direction of transmission (for each cable length):

average = 2.5 percent
maximum = 7.5 percent.

C. Capacitance unbalances, measured at 800 cycles per second, are given in Table II.

D. Mutual impedances between circuits:

Circuits for the same direction of transmission, measured in units of 10^{-9} henry at 5000 cycles: Between side circuits in the same quad—150 average, 540 maximum.

Between pairs of different quads, and between pairs: 115 average, 400 maximum.

Circuits for opposite directions of transmission, measured at 10 000 cycles: 0.9 average, 8 maximum.

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[1] A. B. Clark and B. W. Kendall, "Carrier in Cable," *Bell System Technical Journal*, v. 12, pp. 251–263; July, 1933.

[2] A. S. Angwin and R. A. Mack, "Modern Systems of Multi-Channel Telephony on Cables," *Journal of The Institution of Electrical Engineers*, v. 81, pp. 575–606; November, 1937.

[3] Groningue-Leeuwarden cable, laid in 1935.

1.3 CABLE CHARACTERISTICS AFTER INSTALLATION

Values for one repeater section:

A. Characteristic impedance at 60 kilocycles: ±5 percent difference from the mean value.

B. 60-kilocycle attenuation: Less than 0.2 neper per kilometer.

C. Far-end cross talk between pairs for the same direction of transmission: The lower limits, in nepers (to be maintained in the 12–60-kilocycle band), are as follows—Cross-talk deviation before compensation is 6.3 for all measurements, and 6.9 for 95 percent of all measurements; and cross-talk deviation after compensation is 7.5 for all measurements, and 8.1 for 95 percent of all measurements.

D. Near-end cross talk between pairs of the same direction of transmission: The lower limits of this near-end cross-talk attenuation, expressed in nepers, are as follows—At 30 kilocycles, all measurements should be above 6.5, and 95 percent should be above 6.8. At 40 kilocycles, the values are 6.1 and 6.5, respectively; at 60 kilocycles, 5.8 and 6.1.

E. Near-end cross talk between pairs of opposite directions of transmission: The lower limits of this near-end cross-talk attenuation are 13.8 nepers for all of the measurements, and 14.4 nepers for 95 percent of all measurements.

2. *Design and Manufacture*

The structure of the Paris-Calais and of the Sangatte-Saint Margaret cables is indicated in Table III. The loaded circuits are normal telephone circuits consisting of 0.9-millimeter copper conductors grouped into multiple twin quads with a nominal capacitance of $38.5 \cdot 10^{-9}$ farad per kilometer. This cable also contains pairs (for the transmission of radio-broadcast sound) made of 1.4-millimeter copper conductors. The nonloaded circuits, designed for 12-channel telephone operation, are cabled in star quads and pairs. One group of 7 or 8 star quads, assigned to the same direction of transmission, forms the cable center, over which an electromagnetic shield consisting of alternate layers of copper and steel tape is applied. Over the shield is a group of 16 pairs assigned to the direction of transmission opposite to that of the star quads. The loaded telephone circuits are grouped in the outer layers. Fig. 1 shows the cable cross section (Paris-Clermont section). This is the first cable in which the two transmission directions of one 12-channel link have been included under a common lead sheath; and, so far as is known, it is the only cable of its kind in Europe or in America.

TABLE III

STRUCTURE OF PARIS-CALAIS AND SANGATTE-SAINT MARGARET CABLES

Sections	Length Kilometers	Nonloaded 12-Channel Circuits		Loaded Circuits						Number of Connections
				2-Wire Circuits (Quads)		4-Wire Circuits				
						North-South		South-North		
		North to South	South to North	0.9 mm Copper	1.0 mm Copper	Quads	Broadcasting Pairs	Quads	Broadcasting Pairs	
Paris-Luzarches	33.54	8Q	16P	68	3	11	2	11	2	471
Luzarches-Clermont	33.25	16P	8Q	68	3	11	2	11	2	471
Clermont-Vendeuil	30.9	8Q	16P	40	5	11	2	11	2	393
Vendeuil-Amiens	31.72	16P	8Q	40	5	11	2	11	2	393
Amiens-Beauval	29.72	8Q	16P	31	3	11	2	11	2	360
Beauval-Saint Pol	29.10	16P	8Q	31	3	11	2	11	2	360
Saint Pol-Lillers	24.90	8Q 16P	16P 8Q	40	5	11	2	11	2	393
Lillers-Tilques	35.45	8Q	16P	35	3	9	2	9	2	360
Tilques-Calais	34.41	16P	8Q	35	3	9	2	9	2	360
Calais-Sangatte	9.14	7Q	16P	4	0	12	0	12	0	252
Sangatte-St. Margaret	36.00	16P	7Q	0	0	0	0	0	0	168

Q refers to quads, and P to pairs (12-channel carrier circuits).

The manufacture of the Paris-Calais cable does not present any new aspects; the requirements of 12-channel circuits demand cabling machines (more especially twisting machines) having a precision greater than that necessary for the production of voice-frequency circuits, but does not require the utilization of special machines.

The production of the submarine cable, however, was a delicate operation; to avoid splicing during the laying operations, an effort was made to produce the Sangatte-Saint Margaret cable in only one length. The problem was to manufacture an uninterrupted cable length of approximately 45 kilometers. This work was accomplished in several stages.

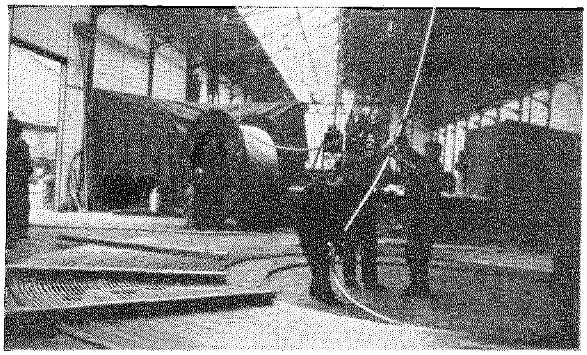

Fig. 2—Coiling of the first manufactured cable length preparatory to splicing; 41 similar lengths were laid over each other as a stack.

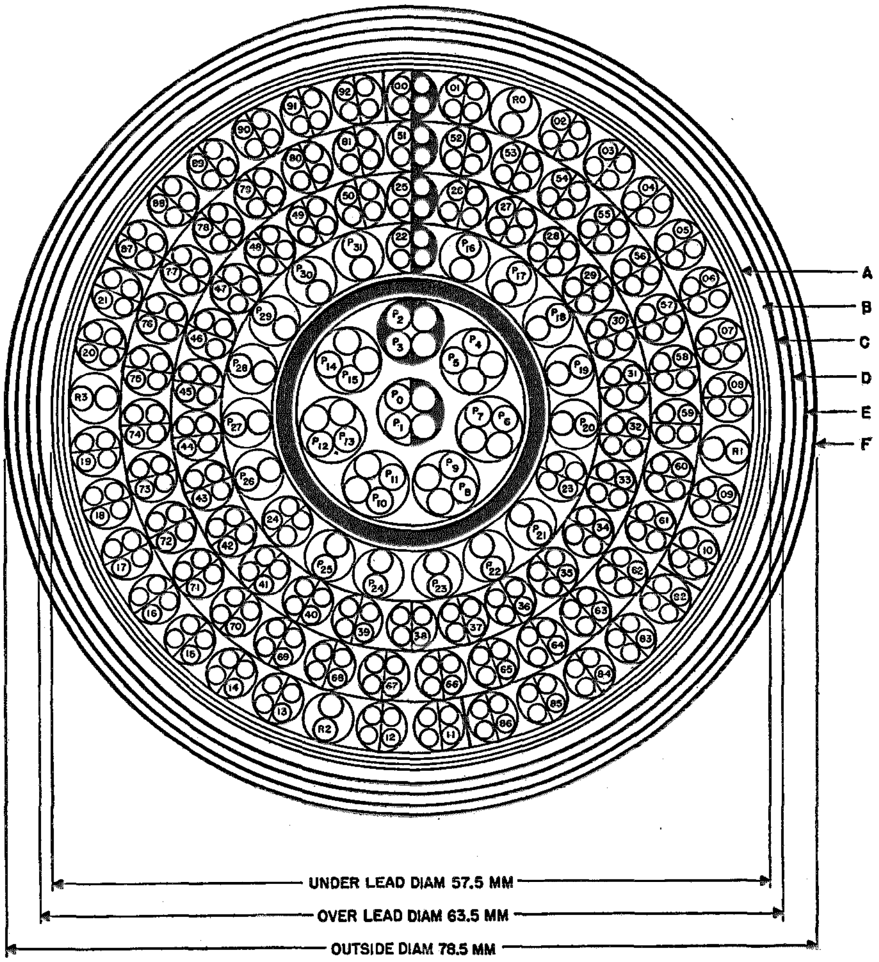

Fig. 1—Cross section of the Paris-Clermont (Oise) cable, consisting of 8 star quads (1.3-millimeter diameter), 16 pairs (1.3 millimeters), 3 DM quads (1 millimeter), 90 DM quads (0.9 millimeter), and 4 broadcast-program pairs (1.4 millimeters). The protective layers consist of: *A*—4 high-voltage paper tapes and 1 cloth tape; *B*—1-percent lead-tin sheath, 3 millimeters thick; *C*—2 impregnated-paper tapes; *D*—2 layers of impregnated jute; *E*—2 painted iron tapes; and *F*—2 layers of impregnated jute.

The cable core, i.e., the 7-quad central group, the electromagnetic shield, the 16-pair group outside the shield, and the paper wrappings were cabled in 400-meter lengths at the Conflans-Sainte-Honorine works of Lignes Télégraphiques et Téléphoniques. The lengths thus prepared were placed under a temporary lead sheath and sent by railroad to the submarine-cable plant of the firm, Les Câbles de Lyon, at Calais. The following operations were then performed: All splices were made; then, in a continuous sequence, the temporary lead sheath was removed, and a continuous lead sheath applied. Finally, the prescribed protective armoring was bound over the sheath.

Fig. 2 shows the coiling of the first cable length on the ground during splicing. The first 41 cable lengths were thus placed over each other and formed a first stack; these coils were so disposed that the ends of all lengths were brought out in the vicinity of two points equipped for the splicing operations and designated (Fig. 3) as stations 1 and 2. The next 41 lengths were disposed in the same manner, so as to form a second stack.

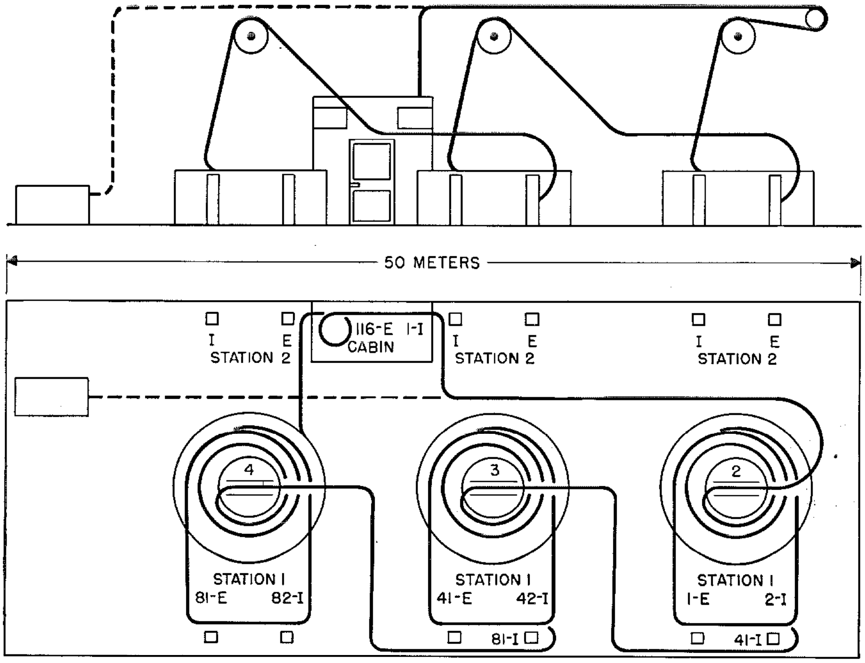

Fig. 3—Arrangement of 3 stacks of cable for splicing at stations 1 and 2. There were 41 lengths of cable in each of the stacks designated 2 and 3 and 34 lengths in 4.

Fig. 4 shows the coiling of the first length of this second stack. Behind it can be seen the stack formed by the first 41 lengths. Finally, a third stack similar to the others, contained the last 34 lengths. At each joining point, the conductor splicings were staggered over approximately two meters of cable so as not to cause any excess over-all thickness; at these points, the circuit twist was carefully preserved, and the continuity of the shield separating the two directions of transmission was restored. An absolute rule was made that there should be no overlapping of quads or pairs, each splicing point being covered with a lead sleeve having the same transverse dimensions as the cable sheath. This sleeve was closed lengthwise with tin solder and its ends were similarly soldered to the lead sheath of the cable.

Starting on December 23, 1938, by February 23, 1939, the cable was entirely spliced and ready for further operations. Transmission tests had shown it to be satisfactory. This cable was approximately 45 000 meters long.

The following operations were performed by the firm, Les Câbles de Lyon. The 45 kilometers of cable were unwound, stripped of the temporary lead sheath, and the final lead sheath was applied. The operations for the whole cable were made in one run and required eight days.

On coming out of the lead press, the cable was coiled inside a tank and the water-tightness test was made. This test was made under a pressure of 1 to 2 kilograms per square centimeter (approximately 15 to 30 pounds per square inch), and, because of the propagation velocity, the pressure test lasted for two months.

Before armoring, the cable was wrapped in a rubber sheath of 6-millimeter thickness. This sheath consisted of 6 vulcanized-rubber tapes. To obtain a uniform and compact sheath, the rubber tapes were covered with an accelerating self-vulcanizing solution. An intermediate tape, removed before applying each rubber tape, prevented the tapes from sticking together.

A layer of jute thread, 3 millimeters thick, was

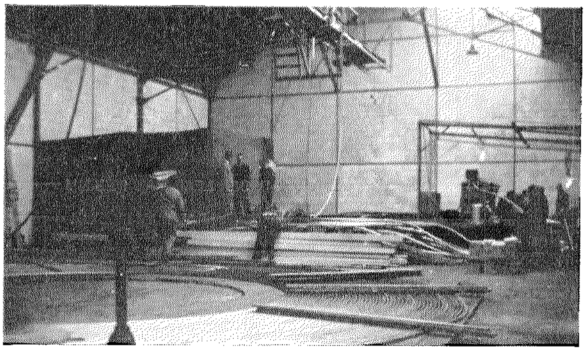

Fig. 4—Starting the second stack, the first stack of 41 cable lengths is in the background.

then applied over the cable. Armor wires, 7 millimeters in diameter, and finally several jute layers soaked with coal-tar and pitch were applied.

The insulation resistance between armor and cables, after a 48-hour immersion, was required to be at least 2 megohms. At the time of laying, it remained above 1000 megohms after several weeks of immersion.

The manufacture of the cable was entirely completed on June 25, 1939. The French Telephone Administration then made complete check tests and found that the cable satisfied the requirements laid down in the specification. The cable was then coiled on board the cable ship, *Ampère*, and was laid on July 25, 1939.

Cable laying in the Strait of Dover is especially difficult because of strong currents flowing between the North Sea and the English Channel, the current direction varying with the tides. This operation was successfully accomplished without serious incident.

On July 26, 1939, it was possible to talk over the circuits between Sangatte and Saint Margaret's Bay. Tests showed the circuit quality to be satisfactory and in December, 1939, this cable was partially used for 12-channel links between Calais and London.

3. *Results*

Some of the properties given in Section 1, such as capacitance, characteristic impedance, and attenuation, are primarily of interest in considering propagation over one circuit. Others, such as capacitance unbalances, mutual impedances, and cross talk, are related to mutual circuit interactions. These two types of conditions will be discussed successively.

3.1 Propagation Over One Circuit

Propagation over one circuit is influenced by the characteristic impedance and attenuation. Figs. 5 and 6 show variation of these factors as a function of frequency for 1.3-millimeter copper conductors grouped in circuits and having a capacitance of $35.0 \cdot 10^{-9}$ farad per kilometer.

Longitudinal uniformity of each circuit is essential to good quality; characteristic-impedance deviations between successive circuit elements cause reflections of the transmitted signals. These reflections, in turn, cause echo phenomena troublesome in themselves and liable to cause an instability of the 2-wire circuits between which is inserted the 4-wire circuit formed by the 12-channel link. Furthermore, they cause near-end cross-talk voltages due to reflected far-end cross-talk voltages and these cross-talk voltages cannot be compensated. This is why a special effort was made to decrease irregularities in the

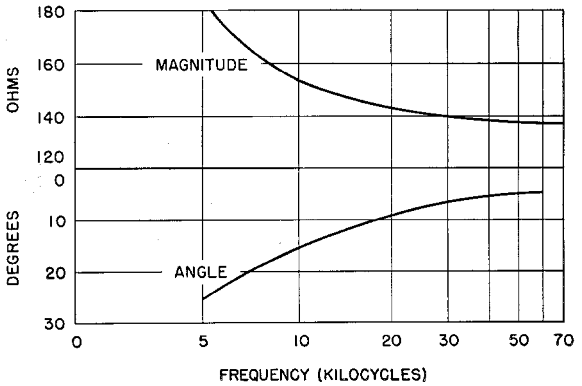

Fig. 5—Characteristic impedance of a nonloaded circuit, the copper conductors are 1.3 millimeters in diameter and their capacitance per kilometer $= 35 \cdot 10^{-9}$ farad.

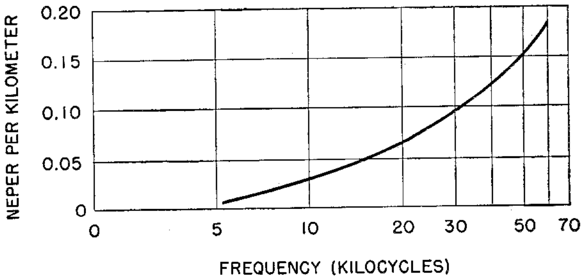

Fig. 6—Attenuation of a nonloaded circuit, the copper conductors are 1.3 millimeters in diameter and their capacitance per kilometer $= 35 \cdot 10^{-9}$ farad.

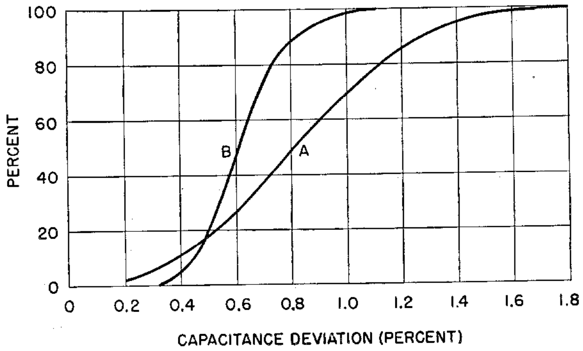

Fig. 7—Average capacitance deviations: *A*—in the 12-channel star-quad group; *B*—in the 12-channel pair group.

characteristic impedance; first by insuring manufacturing uniformity, then by distributing the factory lengths so as to decrease the differences between the average capacitances of successive lengths, and finally, by a special selection at certain junction points of the circuits to be spliced together. As the characteristic impedance of a circuit cannot conveniently be measured directly, the circuit capacitance is generally controlled; capacitance being a basic element of the characteristic impedance.

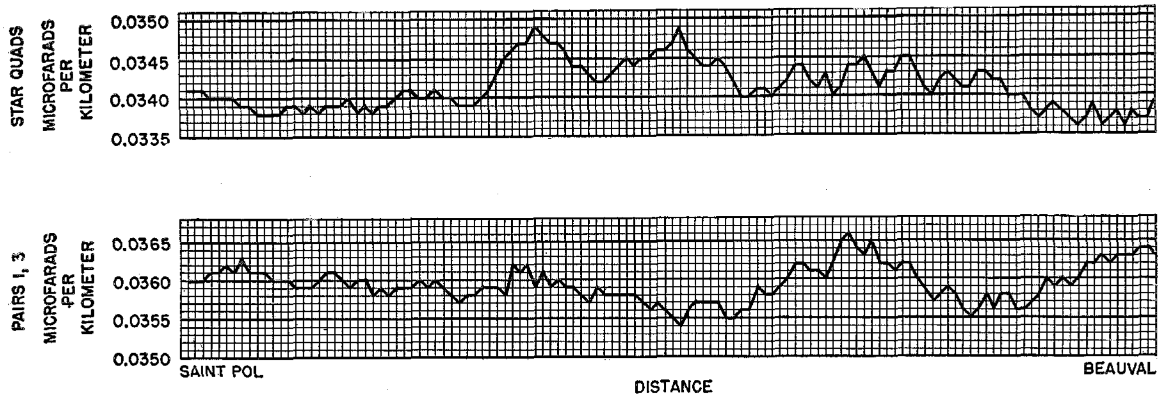

Fig. 8—Carrier-channel circuits, average effective capacitances per length for the section between Saint Pol and Beauval.

Fig. 7 shows a statistical diagram of the capacitance deviations of the factory lengths, i.e., of the differences between the average circuit capacitance and the capacitance of each circuit for each factory length. The ordinates give the percentage of lengths for which the average value of the capacitance deviation is less than the corresponding value shown as abscissa. Fig. 8 shows the average capacitances of the successive lengths of one repeater section. These two diagrams illustrate the uniformity of manufacture.

At the junction points of two successive lengths, three methods of connecting may be used; connecting high-capacitance pairs together and low-capacitance pairs together, or connecting high-capacitance pairs to low-capacitance pairs, or not making any systematic selection. The first method gives different characteristics for the various circuits of the cable, and this may be objectionable. The second method may be used with advantage if the phase variation along a cable length is small enough for the reflection effects on two successive lengths to cancel. It is objectionable if this phase variation is such that the reflection effects on two successive lengths add. Suppose the allocation of the factory lengths has made it possible to make the average capacitance deviations (or the average characteristic-impedance deviations) practically negligible between two successive lengths. With the notations of Fig. 9, which represents the impedance contour for two consecutive lengths AB and BC, we may calculate the resultant reflection coefficient

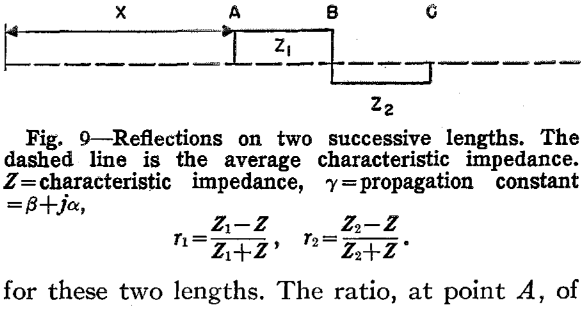

Fig. 9—Reflections on two successive lengths. The dashed line is the average characteristic impedance. Z = characteristic impedance, γ = propagation constant $=\beta+j\alpha$,

$$r_1=\frac{Z_1-Z}{Z_1+Z},\quad r_2=\frac{Z_2-Z}{Z_2+Z}.$$

for these two lengths. The ratio, at point A, of the reflected to the incident voltages is

$$\begin{aligned}R_1&=r_1(1-e^{-2\gamma l})+r_2e^{-2\gamma l}(1-e^{-2\gamma l})\\&=(1-e^{-2\gamma l})(r_1+r_2e^{-2\gamma l}).\end{aligned}$$

If the circuits are selected so that $r_2=-r_1$, the modulus of this expression assumes the value $4r_1 \sin^2 \alpha l$, if the attenuation βl is neglected as compared to the phase variation αl.

If, however, the circuits are spliced without any systematic selection of the capacitances, the modulus of R_1 will vary between an expression close to the above and the value $4r_1 \sin \alpha l \cos \alpha l$. For the condition $r_2=-r_1$ to offer an advantage, it is necessary that $4r_1 \sin^2 \alpha l \ll 4r_1 \sin \alpha l \cos \alpha l$, or $\alpha l \ll \pi/4$. Or, replacing α by its 60-kilocycle value, i.e., 1.8 radians per kilometer, $l \ll 440$ meters.

For the Paris-Calais cable, manufactured in 230-meter lengths, an attempt was made at the junction points of two successive individual lengths to fulfill the condition $r_2 = -r_1$. The advantage thus obtained is probably not very considerable, but it seemed to be worth the attempt. The differences between the input impedances of the various circuits at one end of the cable are shown on Table IV. They show the impedance uniformity obtained.

3.2 Mutual Circuit Interactions

In view of the importance of cross-talk problems, attention should be given to certain fundamental definitions.

3.2.1 Cross-Talk Definitions and Principles

3.2.1.1 Cross-Talk Ratio and Attenuation

Considering two circuits 1 and 2 in the same cable (Fig. 10), the cross-talk ratio E_{12} between these two circuits is the difference between the effective signal level in circuit 2 and the noise level induced in circuit 2 by the signal in circuit 1. The E_{21} cross-talk ratio is not necessarily equal to the E_{12} cross-talk ratio.

The cross-talk attenuation A_{12} between a point P of circuit 1 and a point Q of circuit 2 is the difference between the level at P of the signal in circuit 1 and the level at Q of the noise induced in circuit 2 by the signal in circuit 1.

3.2.1.2 Near-End and Far-End Cross Talk

If a signal is sent out on circuit 1 at the origin O_1 of a cable, the disturbance appearing on circuit 2 at the cable origin O_2 is called near-end cross talk, and the disturbance appearing at the distant end E is called far-end cross talk.

In case these circuits are assigned to the same direction of transmission, the relationship between the cross-talk ratio E_{12} at the distant end and the cross-talk attenuation A_{12} between the origin and the distant end of these circuits is $E_{12} = A_{12} - b_2 + N_2 - N_1$, the factor b_2 designating the attenuation of circuit 2, and N_1 and N_2 representing the signal levels sent out on circuits 1 and 2, respectively.

If the circuits are assigned to opposite transmission directions, this same relationship still exists between cross-talk ratio and cross-talk attenuation at one end of the cable, i.e., at the sending end of circuit 1 and at the receiving end of circuit 2.

For circuits assigned to the same transmission direction, far-end cross talk is the most important, near-end cross talk acting only through the far-end cross talk caused by near-end cross-talk voltages reflected at the origin O_2 of the disturbed circuit. In circuits assigned to opposite transmission directions, near-end cross talk alone has to be considered.

TABLE IV

Impedance Measurements at 60 Kilocycles

Repeater Section	Point Measured	Quad Group Impedance		Pair Group Impedance	
		Average in Ohms	Percent Maximum Difference	Average in Ohms	Percent Maximum Difference
Paris-Luzarches	Paris	142.3	2.2	131.4	1.6
	Luzarches	141.7	1.7	130.6	3.2
Luzarches-Clermont	Luzarches	139.6	1.3	134.6	2.7
	Clermont	141.2	1.1	135.0	1.5
Clermont-Vendeuil	Clermont	142.5	1.0	135.3	1.0
	Vendeuil	140.0	1.7	133.3	1.5
Vendeuil-Amiens	Vendeuil	142.2	1.2	133.7	1.9
	Amiens	144.2	1.7	137.4	1.2

3.2.1.3 Far-End Cross-Talk Couplings

Considering two identical homogeneous circuits 1 and 2, assigned to the same transmission direction and having, at a given point of their route, a coupling M caused by a mutual admittance or impedance, the voltage appearing at the

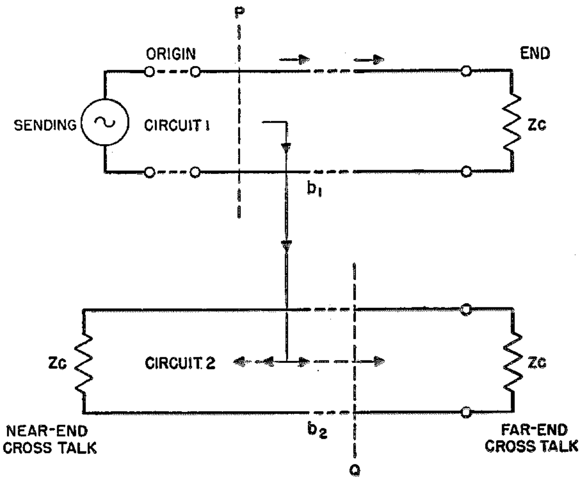

Fig. 10—Near-end and far-end cross talk.

distant end of circuit 2 when circuit 1 is the disturbing circuit (direct cross talk) is the same as that appearing at the distant end of circuit 1 when circuit 2 is the disturbing circuit (inverse cross-talk). This voltage is independent of the location of the coupling *M*. It is important to note that this is no longer true when the two circuits considered are in the presence of other circuits, or when they have impedance nonuniformities, or when they do not have the same transmission characteristics.

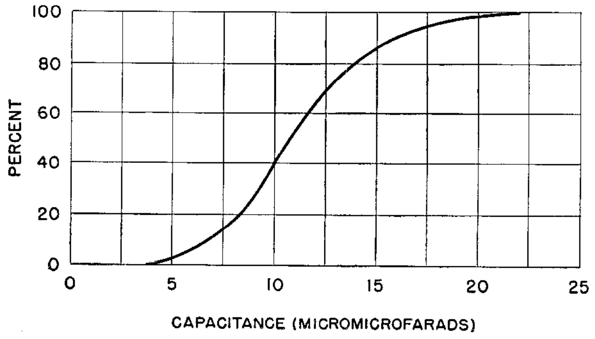

Fig. 11—Real-to-real capacitance unbalances of 12-channel star quads (real circuits of one quad).

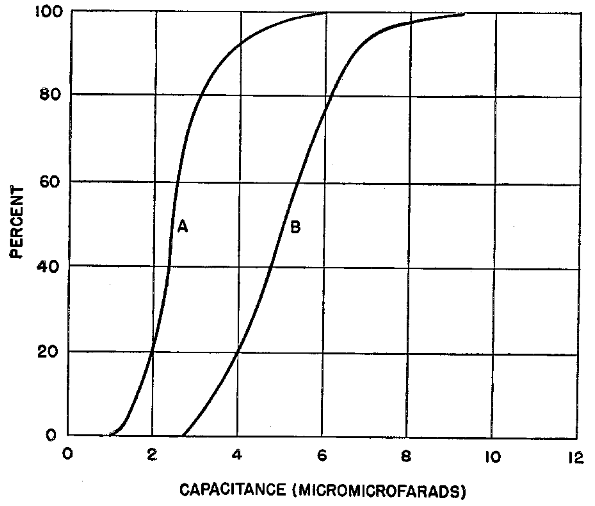

Fig. 12—Real-to-real capacitance unbalances of 12-channel star quads; *A*—between center quad and first-layer quad; *B*—between adjacent first-layer quads.

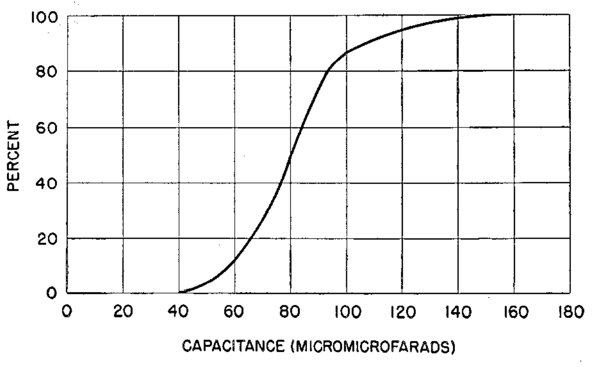

Fig. 13—Real-to-earth capacitance unbalances of the 12-channel star quads.

Accordingly, the resultant equivalent coupling is equal to the algebraic sum of all such elementary couplings distributed along the cable. This is the basis for compensation of far-end cross talk by transpositions at splicing points, and by insertion of a correcting network. It permits a relationship to be established between the average coupling in each shop length and the mean far-end cross talk on a repeater section. As a first approximation, the far-end cross-talk voltage on a cable section is proportional to the square root of the length of the section.

3.2.1.4 Near-End Cross Talk. Segregation of Circuits by Direction

To obtain an adequate cross-talk ratio between circuits of opposite transmission directions, very severe requirements are imposed on interactions between circuits. Consider, for instance, the case of two identical circuits 1 and 2, operated at the same transmitting levels, and the requirement of a maximum cross-talk ratio of 8 nepers at one end of the cable.[4] As the difference between the signal levels on the disturbing and disturbed circuits may be as high as 7 nepers (the maximum attenuation of a repeater section), a minimum cross-talk attenuation of $8+7=15$ nepers between these two circuits at the distant end, permits only an extremely small value of interaction.

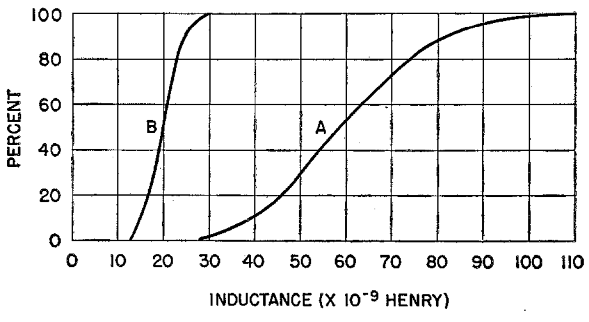

Fig. 14—Real-to-real mutual inductance of 12-channel star quads; *A*—between real circuits of one quad; *B*—between real circuits of different quads.

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[4] C.C.I.F. Document No. 72, Oslo; June 20–July 2, 1938.

This explains the necessity for segregating circuits of opposite transmission directions and for interposing between such circuits a highly effective electromagnetic shield.

3.2.2 Shop-Lengths Problems

The problem consists of manufacturing circuits in such a manner that the mutual admittances and impedances of these circuits be as small as possible. From a practical point, these admittances and impedances involve only capacitance unbalances and mutual inductances.

3.2.2.1 Same Transmission Direction

For circuits cabled in individual pairs, the most favorable lengths of twist must be used. The solution to this problem is made particularly complex by the large number of pairs to be considered simultaneously. For circuits cabled in star quads, it is necessary not only to obtain favorable twist lengths corresponding to small interaction values between pairs of different quads, but also to insure the symmetry of each star quad with a view to decreasing the interactions between the two pairs of one quad. To give concrete figures, the number of pairs of twists to be considered is $8 \times 7/2 = 28$ for interactions between quads, and $16 \times 15/2 = 120$ for interactions between pairs.

The results obtained are shown statistically in Figs. 11 to 17. On these diagrams, the ordinates show the percentages of lengths for which the mean value of capacitance unbalance or of mutual inductance is lower than the corresponding value shown as abscissa. Statistics of capacitance unbalances and mutual inductances between pairs or between side circuits of star quads have been established on lengths of the Clermont-Amiens cable (176 lengths of 223 meters each).

3.2.2.2 Opposite Transmission Directions

To decrease interactions between circuits of opposite directions of transmission, the corresponding twists must first be chosen in an appropriate manner. The number of pairs of twists to be considered is $8 \times 16 = 128$. An appropriate shield must be designed and is an important element of the cable cost, because of its raw materials and of the manpower necessary for its manufacture. It is necessary, therefore, to make a thorough study and to insure maximum shielding.

Theories of shield efficiency have been proposed by Nyquist,[5] Schelkunoff,[6] and Kaden.[7] These theories show the superiority of composite shields consisting of alternate layers of materials

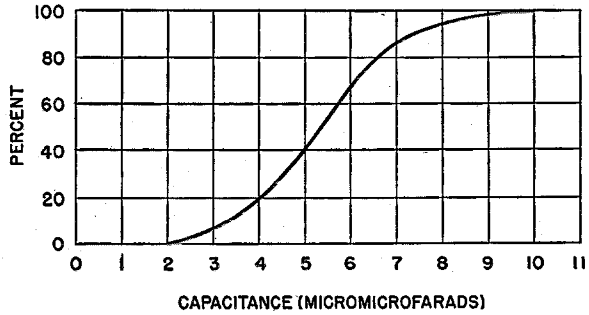

Fig. 15—Capacitance unbalances of two adjacent 12-channel pairs (pair layer).

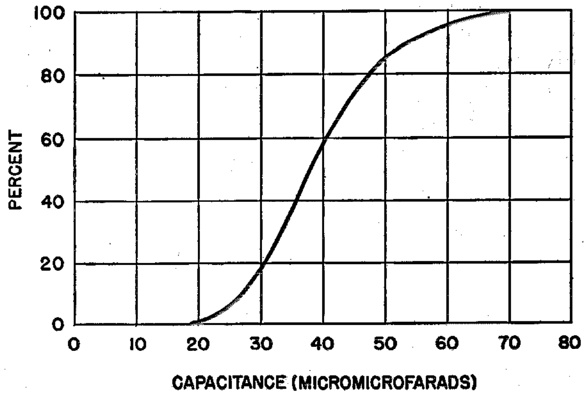

Fig. 16—Ground capacitance unbalances of 12-channel pairs (pair layer).

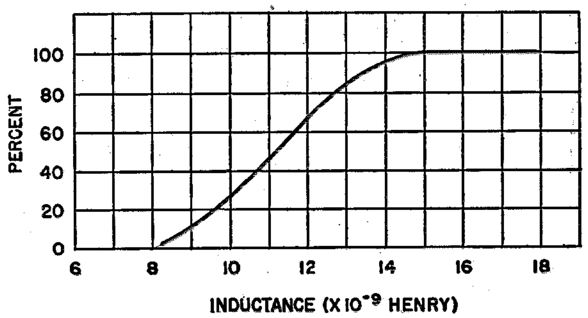

Fig. 17—Mutual inductance between 12-channel pairs (pair layer).

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[5] Nyquist, United States Patent 1 979 402; June 7, 1932.

[6] Schelkunoff, "The Electromagnetic Theory of Coaxial Transmission Lines and Cylindrical Shields," *Bell System Technical Journal*, v. 13, pp. 532–579; October, 1934.

[7] H. Kaden and F. Sommer, "Die Schirmwirkung mehrschichtiger Lagenschirme in Fernmeldekabeln," *Elektrische Nachrichten Technik*, v. 17, pp. 6–16; January, 1940.

having high electrical conductivity and high magnetic permeability. The top and bottom layers should have high electrical conductivity. It is difficult to derive from these theories an accurate value for the efficiency of a shield as practically manufactured, i.e., by means of tape wrappings. Therefore, an experimental determination is of particular interest; Biskeborn[8] indicated a method of measurement involving only short shield samples.

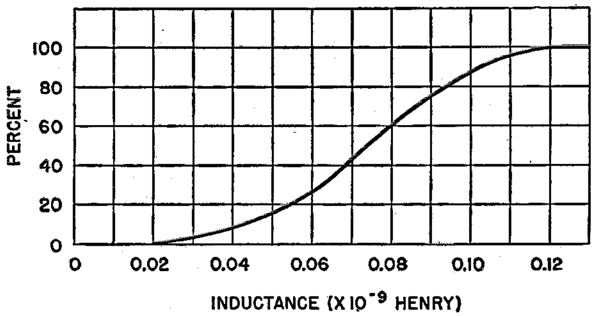

Fig. 18—Mutual inductance between 12-channel pairs and real circuit of star quads (12-channel).

Fig. 18 shows the statistical results obtained for the mutual inductance between circuits on both sides of the shield. These statistics are given in the same manner as above; they were established on lengths of the Paris-Clermont cable (100 lengths of 223 meters each).

3.2.3 Repeater Sections

3.2.3.1 Splicing Lengths. Far-End Cross-Talk Compensation

Sections of eight lengths were first formed and corresponded to loading sections of loaded cables. The eight lengths in each section were spliced in such a manner as to decrease the capacitance unbalances by appropriate circuit transpositions. Then these eight-length sections were connected, two by two, with appropriate transpositions to decrease the differences between direct and inverse far-end cross talk, and, when needed, the high far-end cross-talk couplings.

The sixteen-length sections thus formed were connected, two by two, according to the same principle, and so on until the splicing of a repeater section was complete. It is interesting to compare this method with that used in splicing the first 12-channel cable laid in Sweden.[9]

The results thus obtained on the various repeater sections are shown in Table V. It must be noted that higher values for far-end cross-talk ratios would be needed if compensating elements for far-end cross talk were not inserted at the circuit end. If these compensating elements are not used, an attempt should be made at the splicing points of two cable sections to associate circuits so as to cancel the far-end cross-talk coupling of these two sections. If the insertion of compensating elements is contemplated, an effort is made to remove all factors that limit the improvement caused by these elements: namely the differences between direct and inverse far-end cross talk.

The far-end cross-talk compensation of a repeater section is based on the fact that far-end cross talk at the ends of two identical circuits is independent of the location of the coupling causing it, and remains unchanged when the disturbing and disturbed circuits are interchanged. Thus, far-end cross talk can be compensated by the insertion of coupling elements placed at any point of the circuits, at the ends, for instance. As the characteristic impedance of these circuits is practically a resistance, and as the coupling between circuits is mainly of a capacitive or inductive character, they can all be partly compensated within a wide frequency band by a capacitance unbalance.

A better compensation, however, and practically without any additional cost, is obtained by associating a conductance unbalance with the capacitance unbalance; the compensating element is thus formed by a differential capacitor and one or two resistances connected as shown in Fig. 19. These elements are adjusted so as to obtain a maximum cross-talk ratio in the 12–60-kilocycle band. The values are given in Table V.

The compensation of far-end cross talk between 16 pairs requires the insertion of $16 \times 15/2 = 120$ elements. A relatively convenient mounting of these elements was obtained by passing all circuits through a bare-wire network, a kind of grid in which all circuits are crossed two by two; the balancing differential capacitor is mounted at the crosspoint of two circuits, and the compensating resistors (of the conductance component) are soldered to the differential-capacitor terminals.

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[8] M. C. Biskeborn, "Testing Shields for Carrier-Frequency Line Structures," *Bell Laboratories Record*, v. 18, pp. 88–90; November, 1939.

[9] U. Meyer and W. Rihl, E. F. D., p. 98; July, 1940.

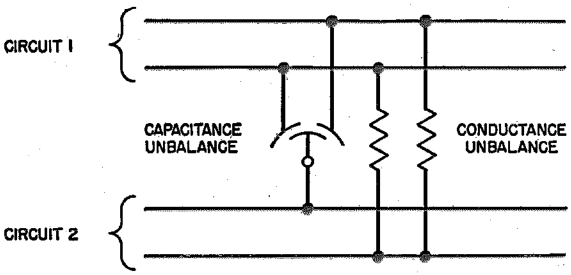

Fig. 19—Complex compensating element for far-end cross talk, inserted between the wires of two circuits.

3.2.3.2 *Submarine Cable*

In the case of the submarine cable, the splicing of the central quad of the 7-quad group presented a special problem. Following a rectilinear path, this quad does not possess the mechanical elasticity of the first-layer quads (cabled along a helix). Thus, in case the cable expands in length, there is danger of this quad breaking. It was therefore decided, in order not to weaken all circuits, to connect together all the central quads of the successive lengths, thus forming a central quad going from one cable end to the other. While this solution offered a mechanical advantage, the central-quad circuits had a propagation constant slightly different from that of the circuits formed by quads of the first layer. For example, the phase difference between these circuits at 60 kilocycles was of the order of 2 percent, or about 80 electrical degrees.

It then became impossible to compensate for direct and inverse cross talk simultaneously by a coupling located at a single point of the circuits. A simple theoretical calculation shows that such a compensation is possible by using two couplings located at two points of the circuits, for instance at the two ends (see Section 4, Appendix). This calculation is confirmed by practical experience.

As an example, the worst 60-kilocycle cross-talk value corresponded to a cross talk expressible by the admittance unbalances: +53.1 micromhos and $174 \cdot 10^{-12}$ farad for direct measurement (cross-talk ratio = 6.5 nepers); and −71.6 micromhos and $105 \cdot 10^{-12}$ farad for inverse measurement (cross-talk ratio = 6.3 nepers).

After compensation, the cross-talk ratio obtained was 9.0 nepers; an improvement of 2.7 nepers.

TABLE V

CROSS-TALK MEASUREMENTS ON 12-CHANNEL CIRCUITS (MINIMUM RESULTS, IN NEPERS)

PARIS-CLERMONT AND CLERMONT-AMIENS SECTIONS

Characteristic	Frequency (Kilocycles)	Paris-Luzarches		Luzarches-Clermont		Clermont-Vendeuil		Vendeuil-Amiens		Specifications	
		Paris Station	Luzarches Station	Luzarches Station	Clermont Station	Clermont Station	Vendeuil Station	Vendeuil Station	Amiens Station	All Measurements	95 Per-cent of All Measurements
Near-end cross talk between circuits of opposite transmission directions	60	15.48	16.20	15.45	16.14	15.12	15.30	15.26	15.34	13.8	14.4
Near-end cross talk between circuits of same transmission direction (high-level end)		P	Q	Q	P	P	Q	Q	P		
	30	8.74	8.36	8.62	9.00	8.84	8.70	8.16	8.64	6.5	6.8
	40	8.46	8.16	7.90	8.42	8.56	8.00	7.96	8.36	6.1	6.5
	60	8.30	7.84	7.46	8.44	8.36	7.70	7.30	8.20	5.8	6.1
Near-end cross talk between circuits of the same direction after compensation (low-level end)		Q	P	P	Q	Q	P	P	Q		
	30	7.62	7.87	7.48	7.78	7.76	8.10	8.16	7.92	6.5	6.8
	40	7.52	7.58	7.34	7.42	7.66	7.74	7.74	7.70	6.1	6.5
	60	7.08	7.36	7.16	7.26	7.44	7.42	7.48	7.46	5.8	6.1
Far-end cross talk between circuits of the same transmission direction before compensation		Q	P	P	Q	Q	P	P	Q		
	60	7.20	7.54	7.32	7.36	7.42	7.38	7.50	7.42	6.3	6.9
	at all frequencies	7.20	7.54	7.32	7.36	7.42	7.38	7.50	7.42	6.3	6.9
Far-end cross talk between circuits of the same transmission direction after compensation		Q	P	P	Q	Q	P	P	Q		
	12	9.46	9.62	9.40	9.60	9.66	9.30	9.16	9.74	7.5	8.1
	30	9.40	9.86	9.50	9.72	9.96	9.46	9.14	10.02	7.5	8.1
	40	9.58	9.74	9.50	9.72	9.94	9.28	9.21	10.06	7.5	8.1
	50	9.50	9.78	9.30	9.66	9.92	9.32	9.08	9.98	7.5	8.1
	60	9.40	9.66	9.32	9.64	9.86	9.30	9.14	9.92	7.5	8.1

P refers to outer pairs of the cable, and Q to the pairs of the central quads.

3.2.3.3 Near-End Cross Talk

Table V gives the near-end cross-talk attenuation obtained between circuits of opposite transmission directions; the equivalent cross-talk ratios are derived from these figures by subtracting the effective attenuations of the circuit, as shown in Table VI.

TABLE VI
EFFECTIVE ATTENUATION OF REPEATER SECTIONS AT 60 KILOCYCLES

Repeater Section	Length (Meters)	Attenuation (Nepers)			
		Per Kilometer		Total	
		Pairs	Quads	Pairs	Quads
Paris-Luzarches	33 540	0.181	0.191	6.06	6.40
Luzarches-Clermont	33 249	0.181	0.192	6.02	6.39
Clermont-Vendeuil	30 905	0.179	0.191	5.54	5.91
Vendeuil-Amiens	31 723	0.178	0.190	5.65	6.03

It will be noticed that the results obtained are appreciably better than the limits specified and that they also exceed the 8-neper cross-talk-ratio limit recommended by the C.C.I.F. (Comité Consultatif International pour La Téléphonie à Grande Distance).

Table V also gives the near-end cross-talk attenuations between circuits of the same transmission direction; the values measured at the "high-level" end are noticeably better than those measured at the "low-level" end. This is due to the fact that the far-end cross-talk-compensating elements are inserted at the low-level end and that their action is superimposed on the circuit near-end cross talk proper.

The construction of the Paris-Calais cable afforded an opportunity for studying the problems presented by the 12-channel circuits, for improving the known solutions, and for applying new solutions. These problems concern more particularly the impedance uniformity of the circuits, the far-end cross talk between circuits having the same direction of transmission, and the near-end cross talk between circuits of opposite transmission directions. The design and manufacture of the Paris-Calais cable has given new experience in 12-channel-circuit technique. This technique is quite varied; it includes special manufacturing methods relative to the twisting and cabling of circuits, the manufacture of an efficient and economical shield, processes for the splicing of lengths in such a manner as to allow an improvement of circuit uniformity and a decrease of coupling between circuits, and adjustments of the far-end cross-talk-compensating elements. It will be noted that the Paris-Calais cable meets not only the requirements laid down by the French Telephone Administration, but also the requirements recommended by the Comité Consultatif International pour La Téléphonie à Grande Distance after the special Paris-Calais cable specification was established. Especially interesting is the cross-talk ratio, the minimum value of which was fixed at 8 nepers for near-end cross talk between circuits of opposite transmission directions and also for far-end cross talk between circuits of the same direction. These values, however, concern only the repeater sections considered separately; the cross-talk disturbances of the various repeater sections add, and it would be particularly interesting to measure all circuits from Paris to Calais, with their intermediate repeaters. Perhaps these transmission tests on 12-channel circuits may be made in the near future; they will form a fundamental basis for the development of new links for 12-channel circuits.

4. *Appendix; Far-End Cross-Talk Compensation Between Circuits Having Different Propagation Constants*

It is known that compensation for far-end cross talk between circuits having similar transmission characteristics can be achieved by placing a complex compensating element at any point along these circuits. If the two circuits have different propagation constants, such compensation is no longer possible and a single compensating element at any point cannot generally balance out simultaneously the far-end cross talk caused on circuit 2 (see Fig. 11 and Paragraph 3.2.1.3) when circuit 1 is the disturbing circuit (direct cross talk), and the far-end cross talk caused in circuit 1 when circuit 2 is the disturbing circuit (inverse cross talk). The calculations given below are relative to the more general case of two circuits with different propagation constants and characteristic impedances. The possibility of compensation by a coupling element inserted at a single appropriately chosen point is first discussed, then, compensation by two coupling elements inserted at two points in the circuits is considered.

Indices 1 and 2 will be assigned to each circuit, respectively; the propagation constant will be designated by $\gamma=\beta+j\alpha$, the characteristic impedance by Z, and the assumption will be made that $\alpha_1>\alpha_2$ and $\beta_1>\beta_2$.

4.1 COMPENSATION AT A SINGLE POINT

Let us place a mutual admittance $ke^{j\varphi}$ at a distance x from the circuit origin. Assuming the same power levels at the origins of the circuits, i.e., a transmission at voltage E_0 on circuit 1 and at voltage $E_0(Z_1/Z_2)^{\frac{1}{2}}$ on circuit 2. If circuit 1 is the disturbing circuit, then:

Voltage at x in circuit 1 is $E_0\, e[\exp -\gamma_1 x]$.
Current at x in circuit 2 is $KE_0\, e[\exp -\gamma_1 x+j\varphi]$.
Current at distant end of circuit 2 is

$$i_{12}=kE_0\, e[\exp -\gamma_1 L] \times e[\exp (\gamma_1-\gamma_2)(L-x)+j\varphi].$$

If circuit 2 is the disturbing circuit:

Current at distant end of circuit is

$$i_{21}=kE_0(Z_2/Z_1)^{\frac{1}{2}}\, e[\exp -\gamma_2 L] \times e[\exp (\gamma_2-\gamma_1)(L-x)+j\varphi].$$

The mutual admittance placed at the ends of the circuits and equivalent to the far-end cross talk may also be measured. Assume $K_1e^{j\phi_1}$ when circuit 1 is the disturbing circuit, causing at the distant end of circuit 2 the current $K_1E_0\, e[\exp\ -\gamma_1 L+j\phi_1]$. Let us assume $K_2e^{j\phi_2}$ when circuit 2 is the disturbing circuit, causing at the distant end of circuit 1 the current

$$K_2E_0(Z_2/Z_1)^{\frac{1}{2}}\, e[\exp\ -\gamma_2 L+j\phi_2].$$

For the mutual admittance $ke^{j\varphi}$ to compensate simultaneously for the direct and inverse far-end cross talk, it is necessary and sufficient that

$$kE_0\, e[\exp -\gamma_1 L]\, e[\exp (\gamma_1-\gamma_2)(L-x)+j\varphi] =K_1E_0\, e[\exp -\gamma_1 L+j\phi_1],$$

and

$$kE_0(Z_2/Z_1)^{\frac{1}{2}} \times e[\exp -\gamma_2 L]\, e[\exp (\gamma_2-\gamma_1)(L-x)+j\varphi] =K_2E_0(Z_2/Z_1)^{\frac{1}{2}}\, e[\exp -\gamma_2 L+j\phi_2],$$

which, by multiplying and dividing, can be brought to the form

$$k^2\, e[\exp 2j\varphi]=K_1K_2\, e[\exp j(\phi_1+\phi_2)], \tag{1}$$

$$\left.\begin{aligned} &e[\exp 2(\gamma_1-\gamma_2)x] \\ &\quad =(K_1/K_2)\, e[\exp 2(\gamma_1-\gamma_2)L-j(\phi_1-\phi_2)].\end{aligned}\right\} \tag{2}$$

The first relationship yields

$$k=(K_1K_2)^{\frac{1}{2}},\quad \varphi=\frac{\phi_1+\phi_2}{2}+\lambda\pi, \tag{3}$$

λ being an integer. The second relationship gives two equations for the determination of x, namely:

$$2(\alpha_1-\alpha_2)x=2(\alpha_1-\alpha_2)L-(\phi_1-\phi_2), \tag{4}$$

$$2(\beta_1-\beta_2)x=2(\beta_1-\beta_2)L-\log_n (K_1/K_2). \tag{5}$$

By eliminating x between these two equations, there is obtained the condition

$$\frac{\log_n K_1-\log_n K_2}{\phi_1-\phi_2}=\frac{\beta_1-\beta_2}{\alpha_1-\alpha_2}. \tag{6}$$

When this condition is fulfilled, the value of x is given by either (4) or (5). To yield a practical solution, this value of x must be such that

$$0<x<L. \tag{7}$$

To discuss (6), an expression is given for the mutual admittances $K_1e^{j\phi_1}$ and $K_2e^{j\phi_2}$. To this end, the total circuit is broken into circuit elements of length dx, along which propagation is neglected, and the mutual admittance designated by $k(x)e^{j\varphi(x)}dx$ is assumed to be constant. Then

$$\left.\begin{aligned} K_1\, e[\exp j\phi_1] &= \int_0^L k(x)\, e[\exp j\varphi(x)] \\ &\qquad \times e[\exp (\gamma_1-\gamma_2)(L-x)]\, dx, \\ K_2\, e[\exp j\phi_2] &= \int_0^L k(x)\, e[\exp j\varphi(x)] \\ &\qquad \times e[\exp (\gamma_2-\gamma_1)(L-x)]\, dx. \end{aligned}\right\} \tag{8}$$

4.1.1 Study of a Few Particular Cases

4.1.1.1 Coupling Uniformly Distributed Along the Circuits

In this case, it can be seen directly that there is equality between the direct and inverse far-end cross-talk currents caused by two elements placed symmetrically with respect to the circuit midpoints. Compensation can then be effected by one additional mutual admittance located at that midpoint.

It can also be seen, starting from (8), that condition (6) is fulfilled, and that (4) and (5) give $x=L/2$.

4.1.1.2 Two Circuits Having Same Attenuation But Different Phase Constants

For $\beta_1=\beta_2$, $\alpha_1\neq\alpha_2$, condition (6) becomes

$$K_1=K_2. \tag{9}$$

This condition being assumed as fulfilled, (4) gives $x=L-\dfrac{\phi_1-\phi_2}{2(\alpha_1-\alpha_2)}$. Condition (7) can then be written

$$0\leqslant L-\frac{\phi_1-\phi_2}{2(\alpha_1-\alpha_2)}\leqslant L,$$

or

$$0\leqslant\frac{\phi_1-\phi_2}{2}\leqslant(\alpha_1-\alpha_2)L. \tag{10}$$

In discussing (9), it should be noted that, practically, the function $k(x)$ varies at random; (8) then shows that the condition $k_1=k_2$ can be fulfilled for certain only if $\varphi(x)$ keeps a constant value φ_0 or $\varphi_0+\pi$. Indeed, in such a case, $e^{j\varphi(x)}$ is equal to $\pm e^{j\varphi_0}$, the + and − signs appearing simultaneously in the two (8) integrals; $e^{j\varphi_0}$ can be taken outside the integral sign, $k(x)$ being positive or negative according to whether the corresponding value of $\varphi(x)$ is equal to φ_0 or to $\varphi_0+\pi$. Besides, due to the fact that $\beta_1=\beta_2$, $\gamma_1-\gamma_2=j(\alpha_1-\alpha_2)$, so that $K_1\,e[\exp j(\phi_1-\varphi_0)]$ is the conjugate of $K_2\,e[\exp j(\phi_2-\varphi_0]$ which involves $K_1=K_2$ and $\phi_1+\phi_2=2\varphi_0$.

In discussing (10), the condition $\phi_1-\phi_2>0$ amounts to fixing a choice on the various possible manners of selecting the angles ϕ_1 and ϕ_2.

In regard to the second inequality of (10), even assuming the hypothesis $\varphi(x)=\varphi_0$ or $\varphi_0+\pi$, the difference between ϕ_1 and ϕ_2 can have any value between 0 and 2π. To fulfill this inequality with certainty, it is therefore necessary that the cable be sufficiently long since

$$(\alpha_1-\alpha_2)L\geqslant\pi. \tag{11}$$

If (11) is not fulfilled, it will be impossible to find a point of compensation in the circuits, when $\phi_1-\phi_2>2(\alpha_1-\alpha_2)L$.

The condition $k_1=k_2$ corresponds to a case very close to practice, when the couplings between the two circuits are limited to capacitance unbalances and to mutual inductances, and when the characteristic impedances of these circuits are real quantities, since the $\phi(x)$ function in the (8) integrals is then equal to $\pm\pi/2$.

4.1.1.3 Two Circuits Having the Same Phase Constant But Different Attenuations

For $\alpha_1=\alpha_2$, $\beta_1\neq\beta_2$, condition (6) becomes

$$\phi_1=\phi_2. \tag{12}$$

This condition being assumed as fulfilled, (5) gives the value of x and condition (7) can finally be written

$$0\leqslant\log_n K_1-\log_n K_2\leqslant 2(\beta_1-\beta_2)L. \tag{13}$$

In regard to (12), as the function $k(x)$ in practice varies at random, condition $\phi_1=\phi_2$ can be insured only if $\varphi(x)$ keeps a constant value φ_0. But this condition is never realized in practice: Suppose the mutual admittance $k(x)\,[\exp j\varphi_0]$ occurs systematically between the circuits considered; there will also occur the mutual admittance $k(x)\,e[\exp j(\varphi_0+\pi)]=-k(x)\,e[\exp j\varphi_0]$, derived from the above by crossing the wires of one circuit. Then, if we require that φ keep a constant value φ_0, this necessitates the splicing of successive circuit elements in such a manner that the couplings add without ever subtracting, which cannot be accepted.

Condition (13) will always be fulfilled as soon as $\varphi(x)=\varphi_0$. But, as shown above, this does not correspond to a practical case. As an example, take the simplified case where the couplings distributed along the cable reduce to two capacitance unbalances, one at the origins of the circuits, $\epsilon k_1 e^{j\pi/2}$, and the other at the distant end, $\epsilon' k_2 e^{j\pi/2}$, ϵ and ϵ' taking indifferently the values ±1. The distant-end mutual admittances (equal to the direct and inverse far-end cross talk) are, respectively,

$$K_1\,e[\exp j\phi_1]=j\{\epsilon k_1\,e[\exp(\beta_1-\beta_2)L]+\epsilon' k_2\},$$

and

$$K_2\,e[\exp j\phi_2]=j\{\epsilon k_1\,e[\exp(\beta_2-\beta_1)L]+\epsilon' k_2\}.$$

In case ϵ and ϵ' have the same sign, it can be seen that (12) and (13) are fulfilled irrespective of the values of k_1 and k_2. In the contrary case, it can be seen that (12) may happen to be fulfilled (but not necessarily), condition (13) being no longer fulfilled.

4.2 TWO-POINT COMPENSATION

In practice, a phase-constant difference is always accompanied by an attenuation difference. Thus, we cannot rely on single-point compensation to produce an appreciable improvement in circuit quality from the standpoint of far-end cross talk. Besides, this method has the drawback that the compensation point is not the same for all circuits in one cable.

It is therefore necessary, if it is desired to effect a systematic and simultaneous compensation of the direct and inverse far-end cross talk, to consider the insertion of balancing elements at several points of the circuits. It can be shown that it is sufficient to locate these elements at two points such as the two ends, for instance.

Let $Ke^{i\varphi}$ be the mutual admittance placed at the near end, and $Je^{i\psi}$ the mutual admittance placed at the far end; the values of K, φ, J, and ψ are defined by the equations

$$\begin{aligned} KE_0\, e[\exp \gamma_2 L + j\varphi] + JE_0\, e[\exp -\gamma_1 L + j\psi] \\ = K_1 E_0\, e[\exp -\gamma_1 L + j\phi_1], \end{aligned}$$

$$\begin{aligned} KE_0(Z_2/Z_1)^{\frac{1}{2}}\, e[\exp -\gamma_1 L + j\varphi] \\ + JE_0(Z_2/Z_1)^{\frac{1}{2}}\, e[\exp -\gamma_2 L + j\psi] \\ = K_2 E_0(Z_2/Z_1)^{\frac{1}{2}}\, e[\exp -\gamma_2 L + j\phi_2]. \end{aligned}$$

In practice, the near-end mutual admittance is first adjusted to compensate for the difference between direct and inverse far-end cross talk. Then the common value of these far-end cross talks is compensated by means of the mutual admittance at the far end.

If the locations of the two compensation points can be chosen, the variation of the compensating voltage as a function of frequency may be made similar to that of the far-end cross-talk voltage to be compensated.

It should be noted that by locating the compensating elements at two points in the circuits, the direct and inverse far-end cross talk will be simultaneously compensated, whatever the cause of the difference between them, provided, however, that the circuit propagation constants be different.

Yarn-Numbering System

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A standard system of numbers for textile yarns would help the electrical industry as well as the textile industry, declares A. A. New, member of the engineering staff of Standard Telephones and Cables Ltd., London, England, in this article.

Mr. New's discussion of the need for a uniform numbering system in Great Britain is especially timely. The American Society for Testing Materials, after several years of study, has just given official approval to such a system—the Grex Universal Yarn Numbering System. This system meets many of the requirements outlined by Mr. New. It is now being discussed with the British Committee on Definitions, which has proposed a similar universal system of numbering.

In the opinion of A. G. Scroggie, chairman of Subcommittee B-2 of ASTM Committee D-13 on Textile Materials, the system is applicable to any type of yarn, including those types used in the insulating of wire and cable. Some of the members of the committee have even suggested that the system might be used for the wire and cable itself, as well as for the insulating yarns, and thus help to bring harmony out of the large number of wire gages currently in use in that industry.

The term "Grex" (plural grex, abbreviation gx) applies to both the unit and the system. It is derived from:

GRams pEr 10,000 meters
GR. . .E. .X

The Grex system meets Mr. New's recommendations in that it is a direct system defined in terms of weight per unit length; it is a metric system; and it permits a large range of magnitudes by the use of multiples or submultiples of the base unit.

THE electrical industry uses large amounts of textile yarns in a wide range of different types of yarn and for this reason suffers particularly from the disadvantages arising from the multiplicity of yarn-numbering systems. It is natural, therefore, that an electrical manufacturing firm should take an interest in adoption of a universal yarn-count system.

The following list shows the number of different yarn sizes typical of those used by a large electrical factory, the total consumption of which may amount to a total of some 1000 tons per annum (excluding fabrics and fabric-base insulating materials): Soft cotton (16 different yarn sizes); flameproofed cotton (2 sizes); mercerized cotton (5 sizes); glazed cotton (4 sizes); cotopa (14 sizes); crestol (5 sizes); natural Chappe silk (1 size); acetate, cuprammonium, fortisan, and nylon (22 sizes); natural silk (Grega selecta) (10 sizes); worsted yarn (2 sizes); linen thread (3 sizes); hempline and seaming twine (3 sizes); jute yarn (2 sizes in Lea count); jute yarn (6 sizes, spindle; pounds per 14400-yard, count).

Of this list of yarns used in a large British electrical factory, the first seven use English cotton count, based on the number of 840-yard hanks weighing 1 pound. However, when used for Chappe silk, the plying numbers have the opposite meaning to that normally employed.

The rayons, Trame, and Grega selecta (but not Chappe) natural silk are numbered in the denier system which is based on the number of half decigrams per 450-meter skein, which is equivalent to grams per 9000 meters. Worsted yarn is supplied on the English worsted count based on the number of 560-yard hanks weighing 1 pound. Linen yarn, linen thread, hempline, and the finer jutes employ the Lea count of the number of 300-yard Leas in 1 pound, while the coarser jutes are denoted by the number of pounds that a spindle of 14400 yards weighs.

Thus we have five systems of counts in regular use, of which two are of direct (or "normal") type (denier count and Dundee jute count) and three of a reciprocal type (English cotton count, English worsted count, and the Lea count). The plying number can have exactly opposite meanings in different cases, and in the case of two [illegible] two different counts are employed for the same kind of yarn in a slightly different state. By a direct (or "normal") type count is meant one in which large count numbers mean coarse, heavy yarns and small numbers mean fine yarns, and by a reciprocal-type count is meant one in which the reverse is the case.

The variety of these different counts affects the work of personnel in the purchasing department, the raw-material inspection department, and the stores and design engineers. The present complexity of what is fundamentally a simple matter in each case causes extra work and increases the possibility of errors.

The effect is noticed chiefly by those concerned with raw-material inspection and with wire and cable design. The ambiguities connected with the methods of expressing the count of multi-ply yarns are also a particular worry of the purchasing department. From the raw-material inspection point of view there are five curious sets of arbitrary constants to be applied in determining counts, where from a common-sense point of view one only would be quite sufficient. This means that a better class of operator has to be used on such work than is really necessary, with a consequent tendency to boredom.

Mechanization of such jobs is difficult or impossible. For a wire or cable, or instrument-cord designer, the multiplicity of count systems and lack of simple relationships makes it more dificult to derive simple fundamental formulae with regard to covering power, diameters, filling power, etc, than would be the case if these were reduced to one system. In addition to these internal effects in an electrical manufacturing unit, there are also external effects of a similar nature if the British firm is concerned with directing or organizing manufacture at a Continental factory. In this case, the complexity of counts is a handicap, although perhaps a minor one, in getting the job done and in getting British yarns into use.

There is, therefore, a strong argument on many grounds for the adoption of a single count system, whatever the system may be. As to which of the many possible count systems should be preferred is a matter of relatively secondary importance. The following points seem worthy of consideration, however:

A. The natural thing is to expect a large number to stand for a large or coarse yarn and a small number for a fine one, or in other words, that the count numbers should vary as a direct function of the size or weight of the yarn; i.e., like the denier system and not as an inverse system like the cotton and worsted counts.

B. Since we are at liberty to choose any numbers, the choice should be the very simplest to calculate and remember, such as the weight per 1, 10, 100, or 1000 units of length.

C. Since the metric system of weights and measures is taught in the schools in Great Britain, is well established in scientific work in that country and for general purposes on the Continent, and export trade is going to be very important in the future, a metric basis seems preferable.

D. A universal system for all textiles includes a very large range of magnitudes. For this reason, the fine silks or cottons at one end of the scale would be hampered by the use of fractional counts, or the coarse jute yarns at the other end would have inordinately large count numbers. This difficulty might be overcome simply by basing the count on the weight of one kilometer of the yarn in grams or kilograms, a "1000" yarn being called a "1K." This would mean that the finest silks and cottons would be about ones or twos while the coarsest jutes (used in electrical work) would become about "30K." Such a system is, of course, merely the present metric denier system with the count numbers divided by 9. It may be argued that this makes the counts of filaments of rayons have numerical values less than 1. This would be avoided by using a 10-kilometer base which would mean that the count numbers would be very close to the present metric denier count (actual 1.1×). This consideration does not seem important, however, compared with the simplicity of the "grams-per-kilometer" basis.

E. With regard to the method of expressing plying, the only logical way is surely that two tens means a twenty. This is a characteristic of the weight-per-unit-length systems and not of the length-per-unit-weight systems with their anomalous principle that "two tens is a five."

To bring about a change to a single simple system would call for some mental readjustment and for the alteration of records and measuring equipment such as wrap reels, yarn scales, etc, but the gain from such a change would be very definite and would be fairly quickly realized.

Behaviour of Telephone Exchange Traffic Where Non-Equivalent-Choice Outlets Are Commoned

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TO give the reader a clear picture of the subject under examination, no better example can be taken than the inevitable result of introducing alternative trunking on inter-exchange circuits. According to theory, the majority of traffic is passed by direct circuits, the remainder being offered an overflow circuit to a tandem exchange, through which the connection can be completed. The number of direct circuits found to be most economical will depend on the type of line construction, the length of the circuit, and similar considerations. It is evident that the overflow traffic offered to the tandem circuit from a number of different groups will be of non-equivalent choice, i.e., some may be the overflow from three direct circuits, some the overflow from four direct circuits, and so on, this overflow traffic being fourth choice, fifth choice, etc.

To determine the proper number of circuits to the tandem centre, it is necessary to establish some basis for equating the relative value of the overflow traffic. A reasonable mean distribution of calls has been deduced by a 50-hour artificial-traffic study. A summary of the results is given in a table and the method for building up the study is described in the appendix.

A theory for equating the traffic on the basis of equivalent first-choice value is explained and compared with the summarised figures of the artificial-traffic study. The probable causes of errors and their extent are examined. This leads to a consideration of the history and unmasking of calls in individual and combined groups. It may be argued that the word "history" is inappropriate, but there are corresponding objections to other terms. The examination considers the case of a combination of two sources of traffic; the identification of the calls that are carried by the first combined choice; and what choice these calls would have been in the two separate groups if they had remained uncombined. The examination also indicates to what extent calls which are masked in the individual group become unmasked in the combined group, and so discloses how the character of the traffic contributes to the efficiency of a group of circuits.

It is not sufficient to ascertain the required number of circuits to the transit centre if the conditions are such that although the over-all loss is accurate, the loss on some portions of the traffic is excessive. This feature is studied by taking practical examples.

A set of tables is included for avoiding the rather eye-straining process of reading off a number of curves to establish equivalent first-choice traffic values. These tables will also be found convenient for other calculations.

The artificial-traffic study deals in units of 2-minute calls, and the summary expresses quantities of 2-minute calls carried by different circuits. The theoretical portion deals with call-hour units because use of this base is more convenient for calculation. It follows that in comparisons it is necessary to transform from one base to the other. Call-hour units are the total number of calls multiplied by the call time expressed in hours.

As far as possible, formulae have been avoided. The reader is invited, nevertheless, to contemplate the formula $a/(1+a)$ which forms a ready means for calculating the traffic deposited on a contact offered a call-hours of first-choice traffic. Frequent use is made of this formula to determine the traffic deposited after the first-choice traffic offered has been ascertained. The formula is derived from the general Erlang formula of loss probability.

This article summarises the conclusions of studies carried out intermittently during the last 15 years. Since these studies were commenced, further examples of cases in which non-equivalent-choice outlets are commoned together have emerged, for example, line-finder systems with partial secondaries, partial outgoing secondaries, etc.

* * *

1. *Illustration of the Case Considered*

The importance of adopting an efficient arrangement of inter-connecting selector outlets has been fully appreciated for many years, and, in consequence, formulae have been developed for determining the number of outlets which are necessary for any particular volume of traffic.

As automatic telephony extends, somewhat abnormal types of inter-connecting selector outlets have been introduced, resulting in new conditions about which little has been published. Some of the artificial-traffic studies which have been undertaken to check the accuracy of various theories thought to be applicable to the new conditions will be summarised.

Most systems of grading outlets assumed that early-choice contacts would be less extensively "commoned" than the later choices.

For example, Fig. 1 shows an arrangement whereby 40 outlets in 4 groups of 10 are commoned to provide 23 circuits. It will be noted that certain fourth-choice outlets are connected together. Similarly, fifth-, sixth-, and seventh-choice outlets are also connected in pairs. All the eighth-, ninth-, and tenth-choice outlets are connected together.

In a commoning arrangement such as that shown in Fig. 2, certain fourth-choice outlets are connected with first-choice outlets. Similarly, fifth and second choices are combined, and so on. The need for providing this type of commoning arises principally with switching arrangements in which there are direct and indirect routes. The scheme is sometimes referred to as a grading with "slipped" commons.

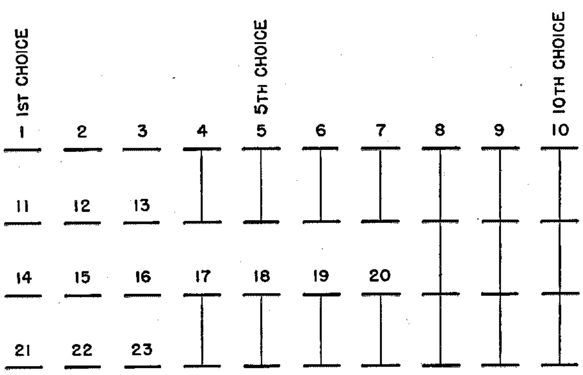

Fig. 1

Methods of computing the number of calls which can be expected to be carried on circuits provided by slipped gradings in which unequal outlets are commoned will be considered.

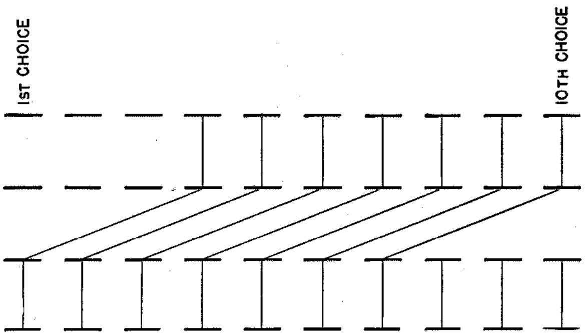

Fig. 2

2. *Analysis of Traffic Values*

To appreciate the general nature of the problem, it is essential to realise that the number of circuits necessary for any particular traffic load cannot be gauged solely by the number of calls of specified duration, as the ability of a circuit to accept such calls also depends on the number of circuits to which the calls have previously been offered. This fact can be illustrated:

If 30 calls of 2-minute duration are offered to two circuits in sequence, the first will accept 15 of the calls (30×5/10), while the second will accept 9 calls (30×3/10), and the remaining 6 will pass over the two circuits. Had the number of calls been smaller, the occupancy of the circuits would have been less, but not proportionally. For example, 15 calls of 2 minutes each would have deposited 10 calls (15×10/15) on the first circuit. An increase of the total calls from 30 to 60 would have resulted in the first circuit taking 20 calls (60×5/15). The figures in brackets illustrate the variation in the ratio of the calls offered to those deposited.

In the first example, 15 of the 30 calls are accepted by the first choice, and the remaining 15 calls are offered to the second choice. The second choice accepts 9 calls. In the second example, 15 calls are offered to the first choice and 10 are accepted.

If a remainder of 15 calls is offered to a third circuit, less than 9 would be accepted, and so on; the later the choice, the smaller being the

TABLE I

48-Hour Artificial-Traffic Study

Hours	Calls Offered	Distribution among Outlets										
		1	2	3	4/1	5/2	6/3	7/4	8/5	9/6	10/7	11/8
Sub-Section 1 = 12 Hours												
2+ 3	132+ 62	40	34	27	9/28	11/12	7/11	3/7	1/3	–/1	–/–	—
8+ 9	111+ 60	43	30	17	6/26	7/17	5/11	2/2	1/2	–/1	–/1	—
10+11	141+ 72	42	33	27	8/30	10/18	6/15	7/5	4/3	2/1	2/–	—
12+13	117+ 65	42	32	22	7/27	7/20	5/12	2/5	–/1	–/–	–/–	—
30+31	115+ 64	37	30	23	10/23	6/20	4/13	3/6	2/2	–/–	–/–	—
50+51	121+ 52	42	31	22	6/26	8/14	6/7	4/4	–/1	1/–	1/–	—
Total	737+375	246	190	138	46/160	49/101	33/69	21/29	8/12	3/3	3/1	—
Mean	[illegible].8+62.5	41	31.7	23	7.7/26.7	8.2/16.8	5.5/11.5	3.5/4.8	1.3/2	0.5/0.5	0.5/0.2	—
Aggregate	185.3	41	31.7	23	34.4	25	17	8.3	3.3	1	0.7	—
Sub-Section 2 = [illegible] Hours												
14[illegible]	133+ 60	42	34	26	14/24	6/21	8/8	3/2	–/4	–/1	–/–	–/–
16[illegible]	126+ 59	40	32	24	6/26	8/19	7/7	5/5	3/1	1/1	–/–	–/–
20[illegible]	109+ 67	37	28	21	8/25	7/16	5/12	2/7	1/3	–/2	–/1	–/1
36[illegible]	128+ 58	43	33	24	8/28	7/16	7/10	5/2	1/1	–/1	–/–	–/–
38[illegible]	121+ 57	40	31	24	10/24	7/15	6/9	3/5	–/4	–/–	–/–	–/–
44[illegible]	119+ 74	40	29	20	7/29	8/18	9/10	4/10	1/4	1/2	–/1	–/–
Total	736+375	242	187	139	53/156	43/105	42/56	22/31	6/17	2/7	–/2	–/1
Mean	122.7+62.5	40.3	31.2	23.2	8.8/26	7.2/17.5	7/9.3	3.7/5.1	1/2.8	0.3/1.2	–/1.2	–/0.2
Aggregate	185.2	40.3	31.2	23.2	34.8	24.7	16.3	8.8	3.8	1.5	0.3	0.2
Sub-Section 3 = 12 Hours												
18+19	114+ 61	42	35	21	5/24	4/18	4/12	1/7	2/–	–/–	–/–	–/–
22+23	126+ 69	42	32	21	8/26	7/18	4/15	4/6	5/2	1/2	1/–	1/–
24+25	125+ 61	40	34	25	10/25	6/19	7/8	2/6	1/1	–/1	–/1	–/–
26+27	121+ 65	41	29	23	9/30	10/18	5/12	3/4	1/1	–/–	–/–	–/–
42+43	120+ 58	39	33	22	7/29	6/18	7/2	2/4	2/3	2/2	–/–	–/–
48+49	130+ 62	41	34	25	7/27	7/18	5/10	4/4	4/1	1/1	1/1	1/–
Total	736+376	245	197	137	46/161	40/109	32/59	16/31	15/8	4/6	2/2	2/–
Mean	122.6+62.7	40.8	32.8	22.8	7.7/26.8	6.6/18.2	5.3/9.8	2.6/5.2	2.5/1.3	0.7/1	0.3/0.3	0.3/–
Aggregate	185.3	40.8	32.8	22.8	34.5	24.8	15.1	7.8	3.8	1.7	0.6	0.3
Sub-Section 4 = 12 Hours												
4+ 5	131+ 62	44	33	22	8/24	9/14	5/13	5/6	3/2	1/2	1/1	–/–
6+ 7	126+ 63	41	29	24	6/28	9/17	9/7	3/6	1/3	2/1	2/–	–/1
28+29	117+ 60	36	32	25	11/24	9/17	2/11	1/5	1/2	–/1	–/–	–/–
34+35	114+ 61	40	33	20	4/30	7/17	5/10	3/3	2/–	–/1	–/–	–/–
40+41	127+ 59	41	33	23	7/24	6/17	6/9	6/4	4/2	1/1	–/1	–/1
46+47	122+ 69	42	33	24	6/28	7/17	6/10	3/6	1/5	–/2	–/1	–/–
Total	737+374	244	193	138	42/158	47/99	33/60	21/30	12/14	4/8	3/3	–/2
Mean	122.8+62.3	40.7	32.2	23	7/26.3	7.8/16.5	5.5/10	3.5/5	2/2.3	0.7/1.2	0.5/0.5	–/0.3
Aggregate	185.1	40.7	32.2	23	33.3	24.3	15.5	8.5	4.3	1.9	1	0.3
Mean												
Sub-Sections 1 and 2	185.25	40.65	31.45	23.1	34.6	24.85	16.65	8.55	3.55	1.25	0.5	0.1
Sub-Sections 3 and 4	185.2	40.75	32.5	22.9	33.9	24.55	15.3	8.15	4.05	1.8	0.8	0.3
Sub-Sections 1–4												
Total	2946+1500	977	767	552	187/635	179/414	140/244	80/121	41/51	13/24	8/8	2/3
Mean (6)	736.5+ 375	244	192	138	47/159	45/103.5	35/61	20/30	10/13	3.3/6	2/2	0.5/0.75
Mean (6)	122.8+ 62.5	40.7	32	23	7.8/26.5	7.45/17.25	5.8/10.2	3.3/5	1.7/2.2	0.6/1	0.3/0.3	0.08/0.13
Aggregate	185.25	40.7	32	23	34.25	24.7	16	8.35	3.8	1.5	0.6	0.2

The distribution of calls is shown with 2-hourly totals because the figures for 1-hourly periods are much more irregular but there appears to be little significance in the greater variation.

Of the 24 2-hourly periods shown, only the hours 26 and 27 are a reasonable representation of the mean. For comparison with the theoretical distribution it would have been preferable to have reduced the number of calls to 120 and 60 so that each group would have been identical.

Traffic data as measured from peg counts or meter readings are frequently the mean, and the 2-hourly totals in the table are a fair representation of this type of variation.

The deviation from the mean appears to indicate that 300 to 400 calls are necessary to provide a mean likely to be within 1 per cent of the true mean. The figures beyond outlet 8/[illegible] can only be considered as very approximate as the number of calls recorded is so small.

acceptance. It is clear, therefore, that if 15 fourth-choice calls are combined with 15 first-choice calls, the combination cannot be considered as 30 fourth-choice calls as this would result in too small an acceptance, whereas if considered as 30 first-choice calls, the acceptance would be too large. Also, if it is determined that the 30 calls combined as described will deposit, say, 12 calls on the outlet to which the calls are offered, will 6 calls come from each group of 15, or will the first-choice calls preponderate, and, if so, to what extent?

3. *Artificial-Traffic Study—Summary of Results*

Table I shows the summarised results of an artificial-traffic study extending over 48 hours. During each hour, the fourth-choice traffic of one group offering approximately 60 2-minute calls per hour is combined with the first-choice traffic of another group offering approximately 30 2-minute calls per hour. The results are presented in four sub-sections, each containing six sets of two hours. The sub-sections have been built up in such a way that the four aggregates are approximately equal and may be compared.

4. *Traffic Calculation Theories*

4.1 APPROXIMATING TRAFFIC BY WEIGHTING METHODS

A rough approximation of the number of circuits necessary for traffic such as that shown in Table I can be obtained by a direct weighting method. The number of calls which may be expected to be deposited on the independent uncommoned outlets may be calculated by established theories. The number of calls that pass to the fourth outlet during each 2-hour period is:

Total traffic, 2-minute calls		122.8
Less: 2-Minute calls deposited on 1st choice,	40.7	
2-Minute calls deposited on 2nd choice,	32	
2-Minute calls deposited on 3rd choice,	23	95.7
Calls reaching 4th choice		27.1

The traffic at the point of combination is

62.5 1st-choice calls, and
27.1 4th-choice calls.

The weighting process involves multiplying the calls by the choice number; the results are added together and [illegible] combined choice number determined. For example:

62.5 (calls) × 1 (1st choice) =	62.5
27.1 (calls) × 4 (4th choice) =	108.4
89.6 (calls)	170.9 (calls multiplied by choice)

Combined choice = 170.9/89.6 = 1.9, or approximately 2, i.e., 2nd choice.

It is known how to determine the exact distribution of 89.6 second-choice calls, but this weighting method suff[illegible]m the disadvantage that the [illegible] be 1.9, or 2.3, or [illegible]ne [illegible]on for [illegible]ch there are no pub[illegible]hed distributions and [illegible]ondly, the division between the fourth-c[illegible] and first-choice sources cannot be estal[illegible]d unless an assumption is made that for [illegible] outlet the division is in the same ratio as [illegible] of the calls offered.

4.2 EQUIVALENT FIRST-CHOICE TRAFFIC THEORY

An alternative means of calculating may be used. This arrangement assumes that there is an equivalent value of first-choice traffic for each number of calls offered to a particular later-choice outlet. In an earlier paragraph, attention is drawn to the fact that in certain circumstances, 15 calls to the second choice deposit 9 calls, while 15 calls offered to the first choice deposit 10 calls. But a number of calls may be offered to the first choice which will deposit 9 calls, and this number may be considered as the equivalent first-choice traffic which can readily be combined with other equivalent first-choice traffic. Furthermore, the ratio between the two groups may be considered on the basis of the equivalent first-choice value.

As an example which may be compared conveniently with the artificial-traffic study, two groups (shown at the base of Table I) have been combined, viz.: 62.5 calls of first-choice traffic (62.5 2-minute calls in 2 hours equal 1.042 call-hours), and that portion of the total of 122.8 calls which passes to the fourth outlet (122.8 2-minute calls in a 2-hour period equal 2.046 call-hours per hour). According to Erlang's theory of probability, the traffic passing to the fourth outlet is 0.452 call-hour.

The combination expressed in call-hours is set out in Table II; the various steps in the calculation being as follows:

A. Taken from column *C* of previous calculation (next line above).

B. Taken from columns *H* and *J* of previous calculation.

C. Column *A* less column *B*.

D. From traffic-offered curves.

E. Column *C* less column *D*.

F. $\frac{\text{column } E}{1-\text{column } E}$ (derived from $\frac{a}{1+a}$ for traffic carried).

G. $\frac{\text{total of column } F}{1+\text{total of column } F}$.

H and *J*. $\frac{\text{individual figures in column } F}{1+\text{total of column } F}$.

5. *Comparison between Theory and Artificial Study*

A comparison may be made between the distribution of calls during a 2-hour period based on:

A. The combined first- and fourth-choice traffic considered as second choice, which is indicated by the weighting process.

B. The groups of late-choice traffic being expressed as equivalent first-choice, and combined together. The figures are taken from the last three columns of Table II, after multiplication by 30 and by 2 to convert from call-hours to 2-minute calls during a 2-hour period.

C. The artificial-traffic study. (See Table I.)

TABLE II

CALCULATION OF EQUIVALENT FIRST-CHOICE TRAFFIC

Traffic to	A Traffic Offered to Previous Outlet	B Traffic Deposited on Previous Outlet	C Traffic Offered to this Outlet	D Hypothetical Traffic to Next Outlet (each sub-group calculated independently)	E Hypothetical Traffic Deposited on this Outlet (each sub-group calculated independently)	F Equivalent First Choice (*E.F.C.*)	G Total Traffic Deposited on this Outlet	H Separate Traffic Deposited on this Outlet	J Separate Traffic Deposited on this Outlet
4th	—	—	0.452	0.213	0.239	0.314	—	0.133	—
1st	—	—	1.042	—	—	1.042	—	—	0.442
						1.356	0.575		
5th	0.452	0.133	0.319	0.139	0.180	0.220	—	0.126	—
2nd	1.042	0.442	0.6	0.255	0.345	0.526	—	—	0.301
						0.746	0.427		
6th	0.319	0.126	0.193	0.082	0.111	0.125	—	0.091	—
3rd	0.6	0.301	0.299	0.104	0.195	0.242	—	—	0.177
						0.367	0.268		
7th	0.193	0.091	0.102	0.038	0.064	0.068	—	0.0587	—
4th	0.299	0.177	0.122	0.038	0.084	0.092	—	—	0.0793
						0.160	0.138		
8th	0.102	0.059	0.043	0.0145	0.0285	0.032	—	0.0301	—
5th	0.122	0.079	0.043	0.0115	0.0315	0.0325	—	—	0.0305
						0.0645	0.0606		
9th	0.043	0.0301	0.0129	0.0038	0.0091	0.0092	—	0.009	—
6th	0.043	0.0305	0.0125	0.0028	0.0097	0.0098	—	—	0.0096
						0.0190	0.0186		
10th	0.0129	0.009	0.0039	0.0011	0.0028	0.0029	—	0.0028	—
7th	0.0125	0.0096	0.0029	0.0005	0.0024	0.0025	—	—	0.0024
						0.0054	0.0052		
11th	0.0039	0.0028	0.0011	0.0002	0.0009	—	—	0.0009	—
8th	0.0029	0.0024	0.0005	0.0001	0.0004	—	—	—	0.0004
							0.0013		

In the case of *A* above, the traffic carried is divided proportionally to the volume offered; in the case of *B*, proportionally to the equivalent first choice.

When comparing these tables (in Table III), the reader should remember that columns *A* and *B* are based on theory so directly that the quantities in each column will form a continuous curve, while those in column *C*, being "observations," are liable to lie above and below a curve which might be formed by an infinite number of observations. Owing to the small value of the later readings in column *C*, it should be appreciated that their significance is minor.

TABLE III

Comparison of Distribution Obtained by Different Methods

Calls offered in a 2-hour period:
1st Choice = 62.5 (1.042 call-hours per hour)
4th Choice = 27.1 (0.452 call-hour per hour)

Choice	*A* Distribution by Weighting Process	*B* Distribution by Equivalent-First-Choice Process	*C* Distribution by Artificial-Traffic Study
1st	23.54	26.52	26.5
4th	10.24	7.98	7.8
Total	33.78	34.50	34.3
2nd	17.48	18.06	17.3
5th	7.60	7.56	7.5
Total	25.08	25.62	24.8
3rd	10.87	10.62	10.2
6th	4.73	5.46	5.8
Total	15.6	16.08	16.0
4th	5.94	4.76	5.0
7th	2.58	3.52	3.35
Total	8.52	8.28	8.35
5th	2.84	1.83	2.1
8th	1.24	1.806	1.7
Total	4.08	3.64	3.8
6th	1.02	0.576	1.0
9th	0.44	0.54	0.5
Total	1.46	1.12	1.5
7th	0.336	0.144	0.3
10th	0.144	0.168	0.3
Total	0.48	0.31	0.6
8th	0.10	0.024	0.13
11th	0.04	0.054	0.08
Total	0.14	0.08	0.21
Overflow	0.04	0.02	—
Total	89.18	89.65	89.56

The small total in Column *A* is attributable to the inaccuracy of the method. A correcting factor could be applied to increase the figure to total 89.6 exactly.

It will be observed that the divergence between column *A* and the other two columns is considerable where the figures are of highest value. The total volume in all columns being equal, it is not unnatural to find that as a result of column *A* being less than column *B* for the first reading, it becomes greater for the fourth, fifth, sixth, seventh, and eighth readings. In these circumstances, the "closeness of fit" in the second and third readings is not very significant.

6. Degree of Error in Equivalent First-Choice Theory

The use of an equivalent first choice is an approximation which is convenient because the traffic carried by an outlet can be easily calculated from any value of offered first-choice traffic. In fact, the equivalent first-choice value is not dependent solely on the volume of the later-choice traffic; it is dependent also, to a very small extent, on the size of the group with which it is combined. The equivalent first choice for 0.5 call-hour offered to the second choice is 0.428571 call-hour. If these two are truly equivalent, two groups of 0.5 call-hour (i.e., 1 call-hour) to the second choice should be equivalent to 2×0.428571 call-hour of first-choice traffic, but it will be seen that there is a small difference. The variation in this difference can be studied in Table IV.

TABLE IV

Traffic Offered to Second Choice	Traffic Carried by Second Choice	Equivalent First Choice Pro Rata on Traffic Offered to Second Choice	Traffic Carried by Equivalent First Choice	Difference
0.5	0.3	0.428571	0.3	0.0
0.761905	0.394089	0.653061	0.39506	+0.00097
1.0	0.460655	0.857142	0.461538	+0.00087
1.3	0.53	1.142857	0.53	0.00
2.0	0.633979	1.714285	0.6315789	−0.0024
3.2	0.7384615	2.742857	0.732825	−0.005636
6.125	0.848077	5.249984	0.8388	−0.009277
0.321428	0.2175825	0.27551	0.216	−0.00158
0.00909	0.00864	0.00779	0.007732	−0.00091

The differences shown in the last column are so small that they are negligible when compared to the inaccuracy of reading curves. As may be expected, the differences are magnified when considering the equivalent first choice for a later choice such as the fourth. See Table V.

In Tables IV and V there is absolutely no significance in the value "0" in the top line of the *Difference* column because the value of equivalent first-choice traffic in this line is chosen so that the traffic carried, as indicated in the second and fourth columns, is equal. In both tables values slightly greater than 0.5 offered call-hour show a positive difference, while much higher and lower values show negative differences. From later tables it would seem that the greatest positive value occurs at the point where the equivalent first choice, when expressed as a percentage of the traffic offered, is at a minimum. To avoid the small error disclosed in Table IV, it is desirable to seek the equivalent first choice for a number of groups as a total, rather than by multiplying the value for a single group by the number of groups. For example, 12 groups each overflow 0.5 call-hour of fourth-choice traffic. This should be calculated as the equivalent first choice to 6 call-hours of fourth-choice traffic (4.69), rather than the individual equivalent first choice, 0.34771 multiplied by 12, or 4.17. When combining traffic values contributed from a number of differing outlets, this inaccuracy can be disregarded; if the traffic total is small, the error is insignificant; if the values are greater, the inaccuracy is swamped by the fact that relative occupancy is calculated on the flat portion of the curve.

TABLE V

Traffic Offered to Fourth Choice	Traffic* Carried by Fourth Choice	Equivalent First Choice Pro Rata on Traffic Offered to Fourth Choice	Traffic Carried by Equivalent First Choice	Difference
0.5	0.258	0.34771	0.258	0.000
0.75	0.34	0.52156	0.34278	+0.00278
1.0	0.411	0.69542	0.41017	−0.00083
1.3	0.482	0.92723	0.48112	−0.00088
2.0	0.585	1.39084	0.58174	−0.00326
3.2	0.707	2.22534	0.68996	−0.01704
6.0	0.825	4.17252	0.80667	−0.01833
0.32	0.187	0.22253	0.182025	−0.004975
0.01	0.0087	0.00695	0.006906	−0.001794

* Read off traffic tables.

In the calculation of the division of traffic in the case taken for the artificial-traffic study, a fourth-choice traffic, rather less than 0.5 call-hour, is combined with approximately 1 call-hour actual first-choice traffic. The equivalent first-choice is seen to give an under-value of 0.00326 (approximate) but, as only 0.34771/1.39084, i.e., 1/4, is equivalent first choice, the probable error is 0.00326/4, or −0.0008, which is still insignificant.

7. *Individual Calls in an Individual Group and in a Combined Group*

The variable behaviour of the pro-rata value of the equivalent first choice raises interest in the history of calls when groups are combined. It seems logical to imagine that certain calls, which would have been deposited on the first choice in individual groups, will be displaced by calls from other groups. Furthermore, it is probable that some calls previously screened will now become deposited on the first choice. Tables VI,

TABLE VI

DISTRIBUTION OF FIRST-CHOICE TRAFFIC WHEN COMBINED WITH FOURTH-CHOICE TRAFFIC

Distribution in Individual Groups	Number of Calls	Distribution in Common Group							
		Outlet 1	2	3	4	5	6	7	8
Outlet 1	383	298	54	20	6	3	1	1	—
2	246	21	160	39	12	10	3	1	1
3	110	6	3	73	21	3	3	1	—
4	41	1	—	3	27	5	3	1	1
5	11	1	—	—	—	5	3	2	—
6	3	—	1	—	—	—	2	—	—
7	1	—	—	—	—	—	—	—	1
	795	327	218	135	66	26	15	5	3

Of the 795 calls, 298 are deposited on the first choice either in an individual or combined group; 21 which are second choice in the individual group become first choice in the combined group; 6 calls are promoted from third choice to first choice; and so on.

TABLE VII

DISTRIBUTION OF OVERFLOW TRAFFIC FROM THIRD OUTLET

Distribution in Individual Groups	Number of Calls	Distribution in Common Group							
		Outlet 1	2	3	4	5	6	7	8
Outlet 4	180	89	55	22	9	4	1	—	—
5	97	7	36	40	6	8	—	—	—
6	42	3	4	12	16	5	1	1	—
7	18	1	—	—	8	5	4	—	—
8	7	—	—	—	—	1	1	5	—
9	2	—	—	—	1	—	—	—	1
	346	100	95	74	40	23	7	6	1

In this case 89 calls are deposited on the fourth choice or on the first common outlet, 7 calls are promoted from fifth choice to first common outlet, and so on.

VII, and VIII show the summarised results of 25 hours with the artificial-traffic study.

TABLE VIII

Percentage Deposited on First Common Choice from Each Source

From	1st-Choice Traffic In Per Cent	4th-Choice Traffic In Per Cent
1st (or 4th)	298÷327=91.13	89
2nd (or 5th)	21÷327= 6.42	7
3rd (or 6th)	6÷327= 1.805	3
4th (or 7th)	1÷327= 0.3	1
5th (or 8th)	1÷327= 0.3	—

The agreement between these figures suggests that there is not any major difference between the distribution of groups of varying origin, and that about 10 per cent is deposited from promoted calls. As would be expected, the last column shows a smaller contribution from the fourth choice and a higher contribution from the later choices.

8. Unmasking Calls in a Group Combination

With some general knowledge as to the probable constitution and history of the calls deposited on a combined outlet produced by a slipped grading, it is possible to study whether an overflow traffic and its equivalent first choice are substantially the same. The theory of the equivalent first-choice traffic for combining different-choice traffic disregards the possibility of any difference in the contribution from calls which fall on other than the first choice when the two groups are uncombined. It is evident, however, that some of these calls are unscreened as a result of the combination. To study the subject, 0.5 call-hour of second-choice traffic may be compared with its equivalent first choice, 0.428571 call-hour. It is natural that the pattern of the second-choice traffic will be more attenuated than the equivalent first choice, and this is clear from Table IX.

TABLE IX

Choice Upon Which Calls Are Deposited	0.5 Call-Hour Second-Choice Traffic	0.428571 Call-Hour First-Choice Traffic
1st		0.3
2nd	0.3	0.104
3rd	0.1375	0.022
4th	0.04712	0.0025
5th	0.01231	—
6th	0.00256	—
Overflow	0.00051	0.000071
Totals	0.5	0.428571

It is to be expected that the contribution from the unscreened calls will be greater from the overflow traffic than from its equivalent first-choice traffic because of the greater number of calls available.

TABLE X

Choice	Calls Offered	E.F.C.	E.F.C. +1	Ratio	Calls Taken
1	2.2	2.2		2.2/3.373	0.653
4	0.148	0.109		0.109/3.373	0.031
21	0.169	0.064		0.064/3.373	0.019
		2.373	3.373		0.703
2	2.2 −0.653 =1.547	1.32		1.32/2.47	0.534
5	0.148 −0.031 =0.117	0.084		0.084/2.47	0.034
22	0.169 −0.019 =0.15	0.065		0.065/2.47	0.026
		1.469	2.47		0.594
3	1.547 −0.534 =1.013	0.77		0.77/1.884	0.4087
6	0.117 −0.034 =0.083	0.0575		0.0575/1.884	0.0305
23	0.15 −0.026 =0.124	0.056		0.056/1.884	0.0297
		0.8835	1.884		0.4689
4	1.013 −0.4087 =0.6043	0.417		0.417/1.496	0.279
7	0.083 −0.0305 =0.0525	0.0368		0.0368/1.496	0.0246
24	0.124 −0.0297 =0.0943	0.0417		0.0417/1.496	0.0279
		0.4955	1.496		0.3315
5	0.6043 −0.279 =0.3253	0.223		0.223/1.27	0.1756
8	0.0525 −0.0246 =0.0279	0.0196		0.0196/1.27	0.0154
25	0.0943 −0.0279 =0.0664	0.0303		0.0303/1.27	0.0238
		0.2729	1.27		0.2148
6	0.3253 −0.1756 =0.1497	0.0997		0.0997/1.13	0.0882
9	0.0279 −0.0154 =0.0125	0.0105		0.0105/1.13	0.0093
26	0.0664 −0.0238 =0.0426	0.0198		0.0198/1.13	0.0175
		0.13	1.13		0.115
7	0.1497 −0.0882 =0.0615	0.0428		0.0427/1.057	0.0405
10	0.0125 −0.0093 =0.0032	0.0024		0.0024/1.057	0.0023
27	0.0426 −0.0175 =0.0251	0.0119		0.0119/1.057	0.0113
		0.0571	1.057		0.0541
8	0.0615 −0.0405 =0.021	0.0148		0.0148/1.022	0.0145
11	0.0032 −0.0023 =0.0009	0.0007		0.0007/1.022	0.0007
28	0.0251 −0.0113 =0.0138	0.0066		0.0066/1.022	0.0064
		0.0221	1.022		0.0216
9	0.021 −0.0145 =0.0065	0.0048		0.0048/1.009	0.0048
12	0.0009 −0.0007 =0.0002	0.0002		0.0002/1.009	0.0002
29	0.0138 −0.0064 =0.0074	0.0037		0.0037/1.009	0.0037
		0.0087	1.009		0.0087
10	0.0065 −0.0048 =0.0017	Remainder			
13	0.0002 −0.0002 =0.0000				
30	0.0074 −0.0037 =0.0037				
	0.0054				

Total traffic offered = 2.517

Overflow = 0.0054/2.517 = 1 in 500, approximately, and considering each group:

2.2 to 1st choice, loss = 1 in 1000
0.148 to 4th choice, loss = 1 in 1000
0.169 to 21st choice, loss = 1 in 50.

These figures indicate that the combination of late-choice traffic with first-choice traffic is unlikely to have an adverse effect on the first-choice traffic.

9. *Probability of Late-Choice Calls Crowding out Early-Choice Calls*

The approximation using an equivalent first-choice value for determining the behaviour of traffic gradings when groups of traffic are combined can be applied to an example:

2.2 call-hours of 1st-choice traffic
0.148 call-hour of 4th-choice traffic
0.169 call-hour of 21st-choice traffic.

It is probable that the third group, being the overflow from a large traffic group, will be liable to considerable variation. Any increase in the number of calls should not swamp the combined group so that the first-choice traffic bears a high probability of loss. The calculations which follow indicate that a large proportion of the calls in the second group are taken; thereafter the calls in the first group are taken, leaving the majority of the remaining calls in the third group. If the number of calls in the third group were to be increased, it is clear that the other two groups would not be seriously reduced. See Table X.

10. *Calculation of Equivalent First Choice*

Table XI shows the equivalent first-choice traffic for ascending values of traffic when offered to different choices. A further column shows the ratio between the traffic offered and the corresponding equivalent first choice. The ratio figures are read off a series of curves drawn through the plots obtained by dividing the equivalent first-choice figure by the traffic offered to the choice under consideration. Any equivalent first-choice value may be calculated quickly by approximating the ratio in per cent and multiplying this figure by the traffic offered.

TABLE XI

CALCULATION OF EQUIVALENT FIRST CHOICE

Originating Traffic	2nd Choice				3rd Choice				4th Choice			
	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Ratio in %
0.1	0.0091	0.00864	0.0087	95.6	0.00046	0.00043	0.00043	93.5	—	—	—	—
0.25	0.05	0.0439	0.0459	91.8	0.0061	0.0056	0.00563	91.0	—	—	—	—
0.3	0.069	0.0592	0.0628	90.8	0.01	0.009	0.0091	90.0	0.001	0.00092	0.00092	92.0
0.4	0.1143	0.0927	0.1021	89.3	0.0216	0.01876	0.0191	88.6	0.00286	0.00257	0.00258	90.2
0.5	0.167	0.128	0.146	88.3	0.0385	0.0321	0.0331	87.0	0.00633	0.00554	0.00556	87.8
0.75	0.321	0.218	0.279	86.6	0.104	0.079	0.0857	83.1	0.0251	0.0204	0.0208	82.9
1.0	0.5	0.3	0.429	85.8	0.2	0.138	0.16	80.0	0.0625	0.0471	0.0494	79.0
1.25	0.694	0.372	0.592	85.3	0.322	0.201	0.252	78.2	0.1212	0.0845	0.092	75.9
1.5	0.9	0.434	0.767	85.2	0.466	0.264	0.359	77.0	0.201	0.13	0.15	73.7
1.75	1.114	0.488	0.953	85.4	0.626	0.324	0.48	76.5	0.302	0.179	0.218	71.5
2.0	1.333	0.533	1.142	85.7	0.8	0.379	0.61	76.3	0.421	0.23	0.295	70.2
2.25	1.558	0.573	1.341	86.0	0.985	0.429	0.752	76.3	0.556	0.28	0.389	70.0
2.5	1.786	0.606	1.54	86.2	1.179	0.474	0.902	76.5	0.705	0.33	0.492	69.9
2.75	2.017	0.636	1.75	86.5	1.38	0.514	1.06	76.8	0.867	0.377	0.604	69.9
3.0	2.25	0.662	1.96	86.8	1.59	0.55	1.22	77.0	1.04	0.42	0.724	70.0
3.5	2.72	0.705	2.39	87.5	2.02	0.61	1.56	77.7	1.40	0.496	0.984	70.2
4.0	3.2	0.738	2.81	88.1	2.46	0.659	1.93	78.4	1.8	0.56	1.27	70.6
4.5	3.68	0.766	3.27	88.8	2.92	0.7	2.33	79.1	2.22	0.613	1.58	71.2
5.0	4.17	0.788	3.72	89.4	3.38	0.73	2.7	80.0	2.65	0.657	1.92	72.1
5.5	4.65	0.807	4.18	89.9	3.85	0.757	3.12	80.8	3.09	0.693	2.27	73.0
6.0	5.14	0.823	4.66	90.4	4.32	0.779	3.52	81.5	3.54	0.724	2.62	74.0
7.0	6.12	0.848	5.58	91.2	5.28	0.814	4.38	82.9	4.46	0.77	3.35	75.7
8.0	7.11	0.867	6.52	91.8	6.24	0.84	5.25	84.1	5.4	0.81	4.17	77.2
9.0	8.1	0.882	7.48	92.3	7.22	0.86	6.14	85.3	6.36	0.833	5.0	78.7
10.0	9.09	0.894	8.43	92.7	8.2	0.876	7.07	86.4	7.3	0.854	5.85	79.9
11.0	10.08	0.904	9.41	93.1	9.18	0.889	8.01	87.4	8.3	0.87	6.7	81.1
12.0	11.08	0.912	10.36	93.5	10.16	0.899	9.0	88.4	9.3	0.884	7.6	82.1
13.0	12.07	0.919	—	—	11.15	0.908	—	—	10.2	0.895	8.5	83.1
14.0	13.07	0.925	—	—	12.14	0.915	—	—	11.2	0.904	9.4	84.0

TABLE XI—*Continued*

Originating Traffic	5th Choice				6th Choice				7th Choice			
	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Raito in %
0.75	0.0047	0.004	0.00402	85.5	0.0007	0.00061	0.00061	87.1	—	—	—	—
1.0	0.0154	0.0123	0.0125	81.2	0.00307	0.00256	0.00257	83.7	—	—	—	—
1.25	0.0368	0.0277	0.0285	77.4	0.0091	0.00722	0.0073	80.2	0.0019	0.00156	0.00156	82.1
1.5	0.072	0.0507	0.0534	74.3	0.00213	0.01597	0.0162	76.4	0.0053	0.00417	0.0042	79.2
1.75	0.123	0.081	0.088	71.4	0.042	0.0298	0.0307	73.1	0.0122	0.00913	0.0092	76.0
2.0	0.19	0.117	0.133	69.6	0.0734	0.0492	0.0517	70.4	0.0242	0.0173	0.0176	72.7
2.25	0.275	0.1575	0.187	68.0	0.117	0.0741	0.08	68.4	0.043	0.0293	0.0302	70.1
2.5	0.375	0.2	0.25	66.7	0.174	0.1037	0.116	66.6	0.071	0.0456	0.0478	67.4
2.75	0.49	0.244	0.323	65.9	0.245	0.1373	0.159	64.9	0.108	0.0662	0.071	65.4
3.0	0.618	0.288	0.404	65.4	0.33	0.1737	0.21	63.7	0.156	0.09088	0.1	63.8
3.5	0.91	0.372	0.592	64.9	0.539	0.2507	0.335	62.1	0.289	0.15	0.177	61.2
4.0	1.24	0.446	0.804	64.8	0.796	0.3276	0.488	61.3	0.469	0.2177	0.278	59.2
4.5	1.61	0.512	1.05	65.2	1.09	0.3998	0.667	61.2	0.694	0.288	0.404	58.2
5.0	1.99	0.567	1.31	65.9	1.42	0.465	0.87	61.3	0.959	0.3567	0.565	57.8
5.5	2.4	0.615	1.6	66.6	1.78	0.523	1.1	61.7	1.26	0.421	0.727	57.7
6.0	2.82	0.655	1.9	67.3	2.16	0.573	1.34	62.2	1.59	0.479	0.919	57.8
7.0	3.69	0.718	2.55	69.1	2.97	0.654	1.89	63.6	2.32	0.577	1.36	58.6
8.0	4.6	0.765	3.25	70.7	3.83	0.714	2.5	65.2	3.12	0.653	1.87	59.9
9.0	5.52	0.8	4.0	72.4	4.72	0.76	3.16	67.0	3.96	0.71	2.45	61.8
10.0	6.47	0.827	4.78	74.0	5.64	0.794	3.86	68.5	4.85	0.755	3.08	63.6
11.0	7.42	0.848	5.58	75.3	6.57	0.8215	4.6	70.0	5.75	0.789	3.74	65.2
12.0	8.38	0.865	6.4	76.6	7.52	0.843	5.37	71.4	6.67	0.8165	4.45	66.7
13.0	9.35	0.879	7.27	77.7	8.47	0.861	6.19	72.8	7.61	0.8385	5.19	68.2
14.0	10.3	0.891	8.18	78.9	9.43	0.875	7.0	74.1	8.56	0.856	5.95	69.6
15.0	11.3	0.9	9.0	80.0	10.4	0.886	7.78	75.3	9.51	0.871	6.75	71.0
16.0	—	—	—	—	11.4	0.897	8.6	76.4	10.47	0.883	7.55	72.2
17.0	—	—	—	—	12.3	0.905	9.53	77.5	11.4	0.893	8.35	73.6
18.0	—	—	—	—	—	—	—	—	12.4	0.902	9.3	75.0

Originating Traffic	8th Choice				9th Choice				10th Choice			
	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Ratio in %
1.75	0.00303	0.00237	0.00237	78.2	—	—	—	—	—	—	—	—
2.0	0.0069	0.00516	0.0052	75.3	0.00172	0.00134	0.00134	77.9	—	—	—	—
2.25	0.0138	0.0099	0.01	72.6	0.00387	0.0029	0.0029	74.9	—	—	—	—
2.5	0.025	0.0172	0.0175	70.1	0.0078	0.00562	0.00565	72.4	0.00216	0.00162	0.00162	75.0
2.75	0.0418	0.0275	0.0283	67.8	0.0143	0.00993	0.01	70.0	0.00434	0.00316	0.00317	73.0
3.0	0.0656	0.0412	0.043	65.5	0.0244	0.0163	0.0166	67.8	0.0081	0.00568	0.00572	70.0
3.5	0.139	0.079	0.086	62.0	0.0596	0.0366	0.038	63.7	0.023	0.015	0.0152	66.0
4.0	0.251	0.1293	0.1485	59.6	0.122	0.0683	0.073	59.7	0.0534	0.0321	0.0332	62.3
4.5	0.406	0.1885	0.232	57.9	0.217	0.1112	0.125	57.3	0.106	0.0588	0.0625	58.9
5.0	0.603	0.2523	0.341	56.5	0.35	0.163	0.195	55.6	0.187	0.0954	0.104	55.6
5.5	0.839	0.3168	0.464	55.4	0.522	0.2205	0.283	54.1	0.3	0.1405	0.164	52.6
6.0	1.11	0.3791	0.611	55.1	0.73	0.2804	0.39	53.4	0.45	0.192	0.238	51.8
7.0	1.74	0.4903	0.963	55.3	1.25	0.397	0.658	52.6	0.85	0.304	0.436	51.2
8.0	2.47	0.581	1.39	56.3	1.88	0.499	1.0	53.2	1.39	0.412	0.711	51.2
9.0	3.25	0.652	1.87	57.6	2.6	0.584	1.4	54.1	2.02	0.507	1.03	51.4
10.0	4.09	0.707	2.42	59.0	3.38	0.651	1.87	55.3	2.73	0.587	1.42	52.0
11.0	4.96	0.7505	3.0	60.6	4.21	0.705	2.39	56.7	3.5	0.651	1.86	53.3
12.0	5.86	0.785	3.64	62.1	5.07	0.747	2.91	58.0	4.3	0.702	2.36	54.5
13.0	6.77	0.812	4.32	63.7	5.96	0.781	3.51	59.4	5.2	0.743	2.9	55.8
14.0	7.7	0.834	5.03	65.3	6.87	0.808	4.2	61.0	6.1	0.777	3.48	57.0
15.0	8.64	0.852	5.76	66.8	7.79	0.83	4.88	62.7	7.0	0.804	4.1	58.4
16.0	9.59	0.867	6.53	68.1	8.7	0.848	5.58	64.1	7.9	0.826	4.74	60.0
17.0	10.5	0.88	7.3	69.3	9.7	0.863	6.32	65.5	8.8	0.845	5.43	61.7
18.0	11.5	0.89	8.1	70.4	10.6	0.876	7.1	67.0	9.7	0.86	6.13	63.1
19.0	12.5	0.899	8.9	71.5	11.6	0.887	7.8	68.4	10.7	0.873	6.88	64.3
20.0	13.4	0.907	9.74	72.6	12.5	0.896	8.6	68.7	11.6	0.884	7.6	65.5
21.0	—	—	—	—	13.5	0.904	9.5	70.6	12.6	0.894	8.4	66.6
22.0	—	—	—	—	—	—	—	—	13.6	0.902	9.2	67.6

TABLE XI—*Continued*

Originating Traffic	12th Choice				14th Choice				16th Choice			
	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Ratio in %
3.5	0.00256	0.00181	0.00181	70.7	—	—	—	—	—	—	—	—
4.0	0.0077	0.00514	0.00517	67.1	—	—	—	—	—	—	—	—
4.5	0.01925	0.012	0.0121	62.9	0.0025	0.00169	0.00169	67.0	—	—	—	—
5.0	0.0414	0.0242	0.0248	59.9	0.0066	0.00425	0.0043	64.0	—	—	—	—
5.5	0.079	0.0432	0.0452	57.1	0.0152	0.00926	0.00935	61.0	0.00219	0.00144	0.00144	65.0
6.0	0.138	0.0698	0.075	54.2	0.0313	0.0179	0.0182	58.0	0.00535	0.00334	0.00335	62.0
7.0	0.334	0.1445	0.169	50.6	0.101	0.0507	0.0534	53.0	0.0232	0.01308	0.0133	57.3
8.0	0.65	0.239	0.314	48.3	0.245	0.1075	0.1205	49.1	0.073	0.0366	0.038	52.3
9.0	1.09	0.340	0.515	47.2	0.49	0.1855	0.228	46.6	0.179	0.0794	0.086	47.9
10.0	1.63	0.435	0.77	47.2	0.84	0.2752	0.379	45.1	0.365	0.142	0.166	45.4
11.0	2.27	0.519	1.08	47.6	1.3	0.3666	0.58	44.6	0.647	0.2194	0.281	43.6
12.0	2.97	0.59	1.44	48.4	1.86	0.4523	0.83	44.6	1.03	0.3038	0.44	42.3
13.0	3.74	0.649	1.85	49.5	2.49	0.528	1.12	45.0	1.51	0.388	0.63	41.7
14.0	4.54	0.698	2.31	50.8	3.19	0.594	1.46	45.8	2.07	0.466	0.87	42.0
15.0	5.38	0.7365	2.8	52.1	3.95	0.649	1.85	46.7	2.7	0.536	1.15	42.6
16.0	6.25	0.77	3.35	53.5	4.74	0.694	2.27	47.8	3.4	0.596	1.48	43.3
17.0	7.14	0.797	3.93	54.8	5.57	0.7326	2.73	48.9	4.15	0.648	1.84	44.3
18.0	8.04	0.8192	4.53	56.2	6.43	0.7643	3.24	50.3	4.93	0.6915	2.24	45.4
19.0	9.0	0.8378	5.17	57.5	7.3	0.7908	3.78	51.8	5.75	0.728	2.68	46.6
20.0	9.9	0.8536	5.82	58.8	8.2	0.812	4.3	53.1	6.6	0.7593	3.15	47.7
21.0	10.8	0.867	6.5	60.2	9.1	0.832	4.95	54.4	7.46	0.7854	3.65	48.9
22.0	11.8	0.8785	7.2	61.5	10.0	0.8478	5.57	55.6	8.4	0.8076	4.2	50.1
23.0	12.7	0.8884	8.0	62.8	11.0	0.8615	6.2	56.7	9.3	0.8264	4.76	51.3
24.0	13.7	0.897	8.7	63.9	11.9	0.8733	6.9	58.0	10.2	0.8426	5.35	52.5
25.0	14.6	0.9045	9.5	65.0	12.8	0.8835	7.6	59.1	11.1	0.8565	6.0	53.7
26.0	—	—	—	—	13.8	0.8925	8.3	60.1	12.0	0.8686	6.6	54.9
27.0	—	—	—	—	14.75	0.9003	9.0	61.1	13.0	0.879	7.3	56.1
28.0	—	—	—	—	15.7	0.9072	9.8	62.4	13.9	0.8882	7.9	57.3
29.0	—	—	—	—	—	—	—	—	14.9	0.8963	8.6	58.4
30.0	—	—	—	—	—	—	—	—	15.8	0.9035	9.4	59.4

Originating Traffic	18th Choice				20th Choice				22nd Choice			
	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Ratio in %
7.0	0.00418	0.00256	0.00257	61.5	—	—	—	—	—	—	—	—
8.0	0.017	0.00946	0.00915	55.8	0.00318	0.00191	0.00191	60.0	—	—	—	—
9.0	0.0524	0.02626	0.027	51.7	0.0124	0.0068	0.0068	55.0	0.00238	0.00141	0.00141	59.0
10.0	0.129	0.0581	0.0617	47.8	0.037	0.01876	0.0191	51.7	0.0089	0.00485	0.0049	55.0
11.0	0.270	0.1073	0.12	44.3	0.093	0.0422	0.0441	47.4	0.0267	0.01335	0.0135	50.7
12.0	0.49	0.1723	0.208	42.4	0.198	0.0803	0.087	44.0	0.067	0.0305	0.0315	46.9
13.0	0.8	0.2477	0.329	41.1	0.369	0.1335	0.154	41.7	0.144	0.0595	0.063	43.9
14.0	1.21	0.327	0.486	40.2	0.619	0.1988	0.248	40.1	0.275	0.1021	0.114	41.4
15.0	1.7	0.405	0.68	40.0	0.955	0.2715	0.373	39.0	0.473	0.1573	0.187	39.3
16.0	2.27	0.477	0.91	40.1	1.38	0.346	0.53	38.4	0.75	0.223	0.287	38.2
17.0	2.9	0.542	1.18	40.7	1.88	0.419	0.72	38.3	1.1	0.292	0.412	37.5
18.0	3.6	0.598	1.49	41.4	2.45	0.486	0.95	38.5	1.54	0.363	0.57	37.0
19.0	4.3	0.647	1.83	42.3	3.09	0.547	1.2	38.9	2.05	0.431	0.76	37.0
20.0	5.1	0.689	2.22	43.2	3.78	0.6	1.5	39.6	2.63	0.494	0.98	37.2
21.0	5.9	0.725	2.63	44.3	4.51	0.648	1.84	40.5	3.27	0.551	1.23	37.6
22.0	6.8	0.755	3.08	45.4	5.29	0.687	2.19	41.5	3.95	0.602	1.51	38.2
23.0	7.6	0.781	3.55	46.7	6.09	0.721	2.58	42.4	4.7	0.646	1.83	39.0
24.0	8.5	0.803	4.06	48.1	6.9	0.751	3.01	43.5	5.45	0.685	2.18	40.0
25.0	9.4	0.8215	4.6	49.3	7.8	0.776	3.46	44.6	6.25	0.72	2.56	40.9
26.0	10.3	0.8378	5.16	50.5	8.65	0.798	3.95	45.8	7.1	0.747	2.96	41.9
27.0	11.2	0.852	5.75	51.5	9.5	0.817	4.47	47.0	7.9	0.772	3.39	42.9
28.0	12.1	0.864	6.36	52.5	10.4	0.833	5.0	48.0	8.8	0.794	3.87	43.9
29.0	13.1	0.875	7.0	53.4	11.3	0.847	5.55	49.1	9.7	0.813	4.34	44.8
30.0	14.0	0.884	7.6	54.3	12.25	0.86	6.1	50.0	10.6	0.839	4.85	45.7

TABLE XI—*Continued*

Originating Traffic	24th Choice				26th Choice				28th Choice			
	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Ratio in %	Offered	Carried	E.F.C.	Ratio in %
10.0	0.00176	0.00103	0.00103	58.5	—	—	—	—	—	—	—	—
11.0	0.00637	0.00346	0.0035	54.9	—	—	—	—	—	—	—	—
12.0	0.0189	0.00947	0.0095	50.4	0.00454	0.00245	0.00246	53.9	—	—	—	—
13.0	0.0477	0.0219	0.0224	46.9	0.0134	0.00669	0.0067	50.0	0.00322	0.00173	0.00173	53.7
14.0	0.1044	0.0437	0.0438	44.0	0.0339	0.01565	0.0159	46.9	0.00944	0.00472	0.0048	50.8
15.0	0.203	0.0772	0.084	41.2	0.075	0.03194	0.0327	44.0	0.024	0.01114	0.0113	47.1
16.0	0.358	0.1227	0.14	39.1	0.149	0.0578	0.0613	41.1	0.054	0.0231	0.0236	43.7
17.0	0.580	0.179	0.218	37.6	0.269	0.095	0.105	38.8	0.109	0.043	0.045	41.0
18.0	0.878	0.242	0.319	36.5	0.445	0.142	0.165	37.1	0.2	0.0723	0.078	39.0
19.0	1.25	0.31	0.45	35.9	0.689	0.197	0.248	36.0	0.339	0.112	0.126	37.2
20.0	1.7	0.377	0.605	35.5	1.0	0.2605	0.353	35.2	0.536	0.1604	0.191	35.6
21.0	2.216	0.441	0.798	35.4	1.39	0.325	0.48	34.7	0.8	0.2163	0.276	34.5
22.0	2.79	0.501	1.0	35.8	1.85	0.389	0.6361	34.4	1.13	0.277	0.383	33.9
23.0	3.434	0.555	1.25	36.3	2.38	0.45	0.82	34.4	1.53	0.339	0.51	33.5
24.0	4.12	0.603	1.52	36.8	2.96	0.507	1.025	34.6	2.0	0.4	0.67	33.4
25.0	4.85	0.646	1.82	37.5	3.6	0.558	1.26	35.0	2.53	0.458	0.85	33.5
26.0	5.61	0.683	2.16	38.4	4.28	0.605	1.53	35.6	3.11	0.512	1.05	33.8
27.0	6.4	0.716	2.52	39.3	5.0	0.646	1.82	36.4	3.75	0.561	1.28	34.1
28.0	7.22	0.744	2.9	40.2	5.76	0.682	2.14	37.2	4.43	0.606	1.53	34.6
29.0	8.06	0.769	3.3	41.1	6.55	0.714	2.49	38.0	5.15	0.645	1.83	35.2
30.0	8.92	0.79	3.76	42.1	7.36	0.741	2.86	38.9	5.91	0.681	2.12	35.9

10.1 EXAMPLE OF CALCULATION

Combine: 0.2 to 30th choice
1.5 to 12th choice
0.6 to 4th choice.

	Equivalent-First-Choice Value
0.2 (to 30th) should be multiplied by 38.1 =	0.076
1.5 (to 12th) should be multiplied by 47.24 =	0.708
0.6 (to 4th) should be multiplied by 69.9 =	0.419
	1.203

Traffic carried 1.203/2.203 = 0.547 which can be distributed as follows:

From 30th, Calls carried 0.076/2.2 = 0.035, Overflow = 0.165
From 12th, Calls carried 0.708/2.2 = 0.322, Overflow = 1.178
From 4th, Calls carried 0.419/2.2 = 0.19, Overflow = 0.41

Total = 0.547

TABLE XI—*Continued*

Originating Traffic	30th Choice			
	Offered	Carried	E.F.C.	Ratio in %
15.0	0.00664	0.00332	0.00333	50.1
16.0	0.0169	0.0079	0.0080	47.3
17.0	0.0385	0.0167	0.017	44.3
18.0	0.079	0.0317	0.0328	41.5
19.0	0.148	0.0547	0.058	39.1
20.0	0.256	0.0867	0.095	37.2
21.0	0.413	0.128	0.147	35.6
22.0	0.63	0.1772	0.218	34.4
23.0	0.91	0.2324	0.303	33.5
24.0	1.25	0.291	0.41	32.8
25.0	1.67	0.351	0.54	32.3
26.0	2.14	0.409	0.69	32.2
27.0	2.68	0.465	0.87	32.4
28.0	3.26	0.517	1.07	32.8
29.0	3.9	0.564	1.3	33.2
30.0	4.58	0.607	1.54	33.6

11. *Appendix—Artificial-Traffic-Study Procedure*

In conclusion, some notes on the arrangement of the artificial-traffic study may be of interest. Sets of 100 pairs of digits were extracted from the London telephone directory. The pairs of digits chosen were the "hundreds" and "tens" of subscribers' numbers. If adjacent entries were identical they were rejected, as were also all those entries without a "thousand" digit. To obtain a greater chance distribution of digits, it is preferable to modify the pairs of digits as read by adding consecutively 1, 2, 3, 4, 5, etc., to each pair of digits and disregarding any carry over. See Table XII. From the 100 sets of pairs of digits used, all those under 60 were considered to be calls arriving at the minute of the hour indicated by the value of the two digits. The process was then repeated to provide an indication of the number of seconds after the minute for each call. In cases where the minute and second indication for two or more calls were identical, a further two digits were established to indicate the "ten millisecond" period after the coincident minute and second values.

TABLE XII

Digits as Read	Add	Use
14	1	25
62	2	84
49	3	72
80	4	24
28	5	73

The process described above was chosen to give a mean of approximately 60 calls per hour. It seems that for the purpose of measuring the number of calls falling on each of a number of contacts, it would be more satisfactory to choose exactly sixty values, thereby allowing no variation between the hours.

For any other number of calls for the hour period, it is only necessary to vary the quantity selected accordingly.

The calls forming the different groups are then combined in chronological order, the times being shown in different columns, and the outlet numbers being marked in corresponding columns. Prior to starting the first-hour period, a trial 30-minute period is put in to give the first hour a normal start.

It takes about 100 hours to record the distribution of a fifty-hour traffic record incorporating approximately 180 calls per hour, and an equivalent time can easily be spent in the compilation of the statistics so created.

Technical Coordination on an International Basis in Communication and Allied Fields*

By E. M. DELORAINE

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THE New York Section of The Institute of Radio Engineers has heard many valuable papers on technical subjects, but there have been only a few on a problem which is of importance, namely, the degree of international coordination in communication and related systems that is desirable or beneficial to all concerned.

The first question which comes to mind is whether there is a real need for any appreciable effort in this type of coördination; whether, in fact, the problems do not resolve themselves in the course of time without special steps being taken. In the case of communication by wire or cable, one might assume that it is sufficient to specify the transfer conditions at the junction between two systems. The other characteristics of the system might be considered of secondary importance.

In the case of radio transmission, the problem might be considered as even simpler. The characteristics of the transmission medium are fixed for all concerned and both ends must necessarily match the transmission medium. Consequently there is no real problem for interworking.

Such a superficial approach does not correspond to facts, and the past has taught us several lessons in this connection. It has been found in practice that unless agreement is reached, and at the proper time, on a large number of points, the international communication system will not function efficiently or economically. A few examples will illustrate this.

Long-distance communication by wire or cable is being handled more and more by means of carrier systems, either over pairs or coaxial cables. The individual voice bands are transposed to occupy a series of positions in the frequency spectrum. The characteristics of the voice bands and the manner in which they are transposed must be specified in great detail if two systems are to interwork.

A long-distance telephone conversation is preceded, accompanied, or followed by a series of signals; they have to do with the selection of the wanted subscriber, the selection of the toll line, the ringing of the subscriber, the metering of the call on the basis of time and distance, and in some cases with the operation of the automatic toll-ticket printer, etc. If, at the time these various signals were introduced and the methods for their transmission selected, a broad plan had been made for interconnection of large networks on an international basis, it is almost certain that the importance of using only one or very few methods would have been fully appreciated. As no such plan was made, simply because it was not appreciated then that the problem would arise, a great many systems have been introduced with resultant complications.

The problems involved in interworking become more important and more pressing as the complexity of the networks and the number of countries involved increase. The introduction of a multiplicity of uncorrelated systems can affect the system to a point where the automatic operation of a large network, involving many countries, such as the European network at present, and probably the western hemisphere network in the future, may become very difficult. One can say, at least, that the lack of broad planning at the early stages later involves an appreciable amount of additional investment in interworking equipment, and some additional operating costs.

The radio communication situation parallels, to a large extent, the wire and cable network situation. As a matter of fact, in many cases, the radio operating method follows the same pattern as the methods introduced first on wires and cables, bringing, at the same time, all the complexities already indicated. The general introduction of printers on radio circuits is an example involving coordination especially when the service is not terminal-to-terminal, but envisages interconnecting any two subscribers on an international basis.

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The general introduction of single-sideband multichannel international telephone circuits involves problems which are closely parallel to those already mentioned in the case of carrier telephone systems on wire or cable.

In the not-too-distant future, the development of long-distance beamed multichannel ultra-high-frequency telephone circuits with relays will also involve a high degree of international coordination, if these systems are to extend over several countries.

It is not only essential in these problems to recognize the necessity of technical agreements, but it is also equally important to make plans for the probable communication networks of the world, far enough ahead, to recognize those elements which require international agreement.

Some may point out that in many cases the information available is not sufficient to allow making such plans with the required degree of accuracy. While this point must be conceded, it is also true that broad plans of this nature help substantially in visualizing future problems.

An interesting example taken from the past is that of the signals, and their methods of transmission, associated with telephone or telegraph communication, particularly in Europe. Apparently, when the various signals were chosen by the numerous parties interested for selection, ringing, metering, supervision, ticket printing, releasing, and the like, the problems were considered essentially on a regional basis. It was not realized soon enough that, when all networks would have to exchange many of these signals, the lack of coordination at the early stages would raise problems almost impossible to solve without adding much equipment for translating signals and insuring interworking conditions.

What happened in Europe is likely to happen in the world in the future as the scale of distances rapidly decreases, particularly with the introduction of fast air travel.

Large regions of the world are without adequate communication facilities and development of these facilities on a regional basis would later reproduce on a larger scale what happened in prewar Europe, for instance. The air traffic extending over all the world will create new exchanges of goods and commodities and thereby will call for additional communication facilities, following generally the pattern of the airways themselves. In consequence, plans of the probable world communication network in the less-developed countries of the world in relation to air transport appear to be very timely.

Let us consider now, methods which are capable of dealing with the problems just described.

Conferences were called and agreements were reached on certain communication matters before World War I, especially in relation to safety of life at sea. It was not, however, until after World War I that, due to a recognition of the great expansion of communication which was going to take place, the principal international committees were created dealing first with telephony, then telegraphy and radio.

These steps were in a large part the direct result of the presidential address of Sir Frank Gill to the Institution of Electrical Engineers in London in 1922. His proposals for international coordination in communication received a quick response and the Comité Consultatif International pour La Téléphonie à Grande Distance (CCIF) was formed and did exceedingly good work during the period between the two wars. The pattern set by this first committee was followed to a large extent in the creation of two other committees for telegraph and radio, known respectively as Comité Consultatif International Télégraphique (CCIT) and Comité Consultatif International Radio (CCIR).

These committees are, of course, a grouping of the operating agencies. They appoint technical commissions which study the problems listed by the committee. These commissions are permanent. They include operating, manufacturing, and laboratory experts, and they issue information and recommendations on the problems submitted. The recommendations are adopted or rejected by plenary assemblies of the operating agencies. These assemblies take place at intervals of one or two years. The recommendations when adopted are issued as international recommendations. They are widely distributed, and it is a fact that the operating authorities almost invariably specify when procuring equipments that all materials delivered will have to meet the CCI recommendations.

The real difficulties encountered at these meetings are not generally of a technical nature, but more often are due to the fact that a group of nationals may be reluctant to give up some temporary advantage for a general benefit or subordinate any portion of the sovereign rights of their country to matters of international character. Sir Frank Gill, who, since the creation of these committees has consistently been active in these matters, expresses himself in a recent letter as follows:

> I think an illustration of this difficulty of sovereignty may even lie in the sphere of economics. It is conceivable that each of several different countries might find it economical to employ somewhat different systems of communication such that the whole would not give satisfactory results when put together. They must, therefore, find a compromise in economics to get the technical solution required by means of the best *joint* economics they can devise. In other words, the engineers must in such cases cease to think of themselves as working for any particular nation, they must work for the whole group.
>
> I think it is along this line that the CCIF, CCIT, and probably the CCIR (though I am not so familiar with the latter) have done good work. They have faced the difficult subject of international standardization by concentrating on the essential clauses in specifications and they have not attempted to standardize rigidly any piece of apparatus. In the absence of any powers to compel nations to adopt their advice, they have worked in a very splendid manner along the lines of goodwill and although very much remains to be done, they have made very great progress in international (telephone) communication.

In spite of the shortcomings which we have stated, a large degree of order was introduced in Europe and in the world, prewar, due to the work of these organizations. The last war unfortunately partly destroyed this order through the introduction, principally by the Germans, of many non-CCI systems during the war period.

Experience was accumulated prewar on these very important matters. Consideration should now be given to the situation before us to see whether most questions requiring international agreements in our particular field are already solved, or have reached a stage where no further progress in standardization can be made, or whether there is at this stage, a possibility of great development requiring, in turn, that many steps be taken at the present time to prevent confusion in the future.

The latter appears to be the case and results from the many new aspects connected with the development of communication with mobile and air transport. This example of air navigation can be translated to an extent also in terms of telephone communication with cars, trucks, busses, or railroad cars. This latter new facility does not involve problems by far as complicated as aviation, but it does call also for a large degree of coordination. A person interested in telephoning from his car might be surprised to find, in ten years, that he must buy or rent a multiplicity of equipments or adaptors to maintain contacts with the telephone network even over a limited area in the United States, not to speak of the difficulties he might run into if he attempts to cross the borders.

H. M. Pease, Chairman of the Board of International Standard Electric Corporation and a veteran in international communication, says in a memorandum:

> I believe that the main communication arteries in future decades will follow the air routes and that the planning engineer can play an important part in bringing about standardization of systems and apparatus used by the various operating agencies throughout the world.

It is my opinion, after some study of the problem, that the questions of ground communication along the airways, for airways operation or for public use, communication with aircraft, weather reporting, air navigation, airport traffic control, and instrument approach are all interrelated.

These methods cannot be selected on a national basis either, as airplanes already travel from country to country and will do so more and more as time goes on. In consequence, the problems must be considered on a world basis. This requires an elaboration and study of typical world plans of air traffic. These plans cannot be accurate at this stage, but they will surely help to visualize the problems to be solved.

These problems when clearly understood and listed must be studied by the appropriate technical commissions including operating, manufacturing, and research people. These technical commissions should operate very much in the same manner already found so useful in the CCI committees. They would issue their recommendations to be approved by an assembly of operating agencies and these recommendations would in all probability become the basis of future procurement by all airlines.

If this is not done, our effort will be in vain; air transport will be developed on the basis of a multiplicity of systems and methods involving as an immediate consequence a multiplicity of types of equipment, both on the ground and in the airplane. This will cause extra investment, extra load, unnecessary expenses, and a slowing down of the development in the whole field.

Imagine for a moment the complexities that will inevitably result from a lack of agreement in the assignment of radio channels or in the transmission characteristics of the numerous services already mentioned. Imagine, for instance, airports on the north Atlantic lanes, or in South America, or in Africa handling traffic of airlines of ten different nationalities, each differently equipped and utilizing different characteristics for their long-range and short-range communication, for navigation, radio altimeters, distance indicators, apparatus for anticollision or the equivalent, for airport approach, instrument landing, or weather reporting. It makes a rather discouraging picture by its very complexity and one realizes that the problems involved in handling such traffic would be almost impossible to solve.

It is clear, in consequence, that plans must be developed, agreed to, and followed, at least in our particular field, to permit orderly development of air transport.

International organizations already exist for making this study and preparing such plans. The most important bodies are the International Air Transport Association, working principally in terms of requirements, and the Provisional International Civil Aviation Organization, which already has issued some general international recommendations. It is possible however that the importance of the task has not been recognized except by those directly involved, and we may encounter all the difficulties set forth by Sir Frank Gill in the above quotation.

The various technical societies, such as The Institute of Radio Engineers, can help in stressing the importance of technical agreements in these fields.

The objection one can raise to forward planning and standardization is that it presumes knowing the solution to many technical problems yet to be solved. Also, that the advantages of standardization are partly balanced by its tendency to retard technical progress. There is, of course, some truth in both points. A broad plan, however, can be prepared usefully, even with the present knowledge, or in some cases an extension of this knowledge into the immediate probable future. Such plans must not be crystallized but must change as new technical information becomes available. Regarding the second objection, it is necessary to introduce a large amount of free wheeling between development and operation, permitting development to proceed while the operating standard remains unchanged for periods.

The engineers, in general, and those of this Institute in particular, have to play their part in the development of world plans in communication and allied systems to bring about the full benefits which can be derived from the technical progress resulting in large part from their own labors.

Extension of Norton's Method of Impedance Transformation to Band-Pass Filters

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SOME applications of a method of network transformation first discovered by E. L. Norton[1] and sometimes used in the design of band-pass filters are considered. Norton's method can be extended in different ways, and indicates in certain cases the design of new and more economical structures for composite band-pass filters.

• • •

1. *Norton's Method*

One alternative of Norton's method is based on the equivalence between the networks of Figs. 1*A* and *B*, where the transformer is assumed to be ideal. Another alternative can be deduced from the first case by the principle of duality.

In the network of Fig. 1*B*, at least one of the elements is negative, and thus physically unrealizable. The transformation can, however, be applied to a part of a network if the negative impedance can be absorbed into some positive impedance of the remaining part of the network.

For a positive transformation ratio, the series impedance in Fig. 1*B* is positive and only one of the shunt arms is negative. Physical realization requires that the negative arm should be combined in parallel with a smaller positive impedance. In the limiting case, the negative and positive shunt impedances may cancel each other and the result is an open circuit.

This equivalence was used by Norton to incorporate ideal transformers in ladder networks in which one shunt arm and the next series arm are simple elements of the same kind, e.g., both inductances or both capacitances. This occurs in some band-pass filters, and the design of such impedance-transforming filters is described in most textbooks.[2]

2. *Transformation of a π Network Consisting of Antiresonant Circuits*

In the equivalence between Figs. 1*A* and *B*, the impedance Z is not required to be a single inductance or a single capacitance, but may be, for instance, an antiresonant circuit. As a parallel combination of antiresonant circuits is also an antiresonant circuit, the use of this method does not introduce new elements when applied to ladder networks composed of antiresonant circuits and terminated at midshunt.

The simplest case is a single π network (Fig. 2). This network will be modified by replacing the series arm with a network associated with an ideal transformer of arbitrary ratio as in Fig. 1*B*. The transformer can then be suppressed if all the impedances on the right-hand side and the output impedance are divided by k^2. By combining adjacent arms in parallel, a final structure

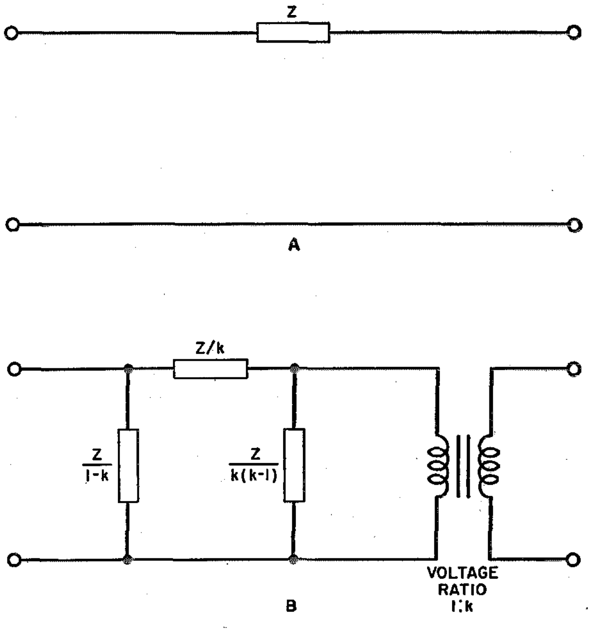

Fig. 1—Equivalent networks.

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[1] United States Patent 1681554.

[2] T. E. Shea, "Transmission Networks and Wave Filters," D. Van Nostrand Company, New York, New York, 1929; pp. 325–334.

is obtained which is obviously a π network similar to the original (Fig. 2). Both networks are electrically equivalent except for the transformed output impedance.

The values of the new circuit elements are easily computed from Fig. 1*B*, and the range of possible values for k is found to be restricted by the condition that all inductances and capacitances have positive values.

Before writing down these conditions explicitly, a general statement will be made regarding physical realizability. The following theorems are easily proved:

A. The frequency of antiresonance of the parallel combination of two physical antiresonant circuits always lies between their individual frequencies of antiresonance.

B. If a physical antiresonant circuit be considered as a parallel combination of two physical antiresonant circuits and one of these circuits be removed, the frequency of antiresonance of the remaining part, as compared to the frequency of the entire circuit, is either *increased* or *decreased*, depending on whether the frequency of antiresonance of the part removed is *lower* or *higher* than that of the original circuit.

Returning now to the transformation of the network of Fig. 2, suppose that k is less than unity so that the output impedance will be increased (in the opposite case, simply replace k by $1/k$ and interchange input and output). The only negative impedance in Fig. 1*B* is thus the shunt arm on the right-hand side, and the only limiting conditions for k are the positive character of the new values L_3' and C_3' which arise from the parallel combination of the original $L_3\,C_3$ circuit with a negative circuit resonating at the same frequency as $L_2\,C_2$. Denote these respective frequencies by f_3 and f_2. To recognize whether the positive character of either L_3' or C_3' is more restrictive, apply the last theorem. If $f_2>f_3$, the combination of the positive and of the negative circuit is equivalent to the removal of a positive circuit with higher antiresonance frequency; the resulting frequency f_3' is thus decreased with respect to f_3 and can, in the limiting case, be reduced to zero. The new circuit is then reduced to a pure capacitance and a further decrease is impossible unless L_3' becomes negative. In this case the most severe restriction arises from the inductance. The opposite result occurs if $f_3>f_2$.

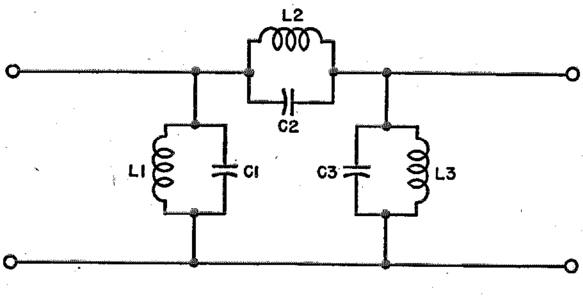

Fig. 2—Single π network using antiresonant circuits as elements.

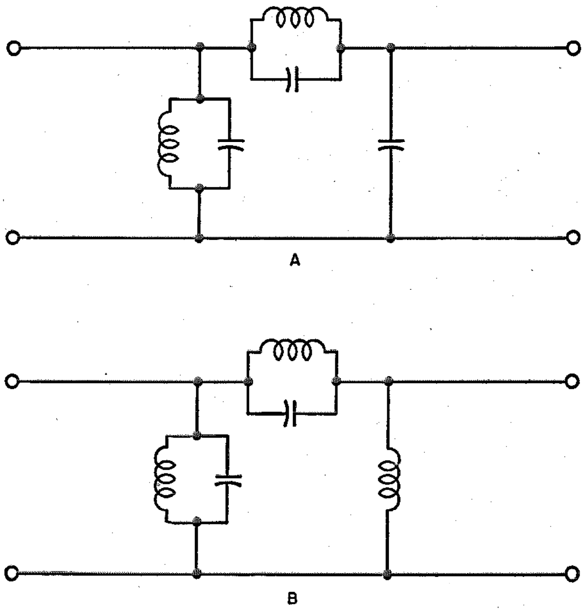

Fig. 3—Reduction of circuit of Fig. 2; *A* for cases where $f_2>f_3$, and *B* if $f_2<f_3$.

In conclusion, the network of Fig. 2 may be reduced to the form of Fig. 3*A* if $f_2>f_3$, or of Fig. 3*B* if $f_2<f_3$, by a proper choice of the transformation ratio. In either case this transformation increases the output impedance as compared to Fig. 2.

An elementary calculation shows that in the case of Fig. 3*A* the ratio should be

$$k=\frac{L_3}{L_2+L_3},$$

and in the case of *Fig.* 3*B*,

$$k=\frac{C_2}{C_3+C_2}.$$

By duality, inverse results are obtained for a T network composed of series-resonant circuits.

3. General Transformation

Consider a ladder network composed exclusively of antiresonant circuits (Fig. 4). This network will be modified by replacing each series arm by a π network associated with a transformer of arbitrary ratio, as in Fig. 1. All the transformers can be omitted if each impedance (and the output impedance) is divided by the square of the product of the ratios of all the preceding transformers. After combining adjacent arms in parallel, a final structure is obtained, which is obviously similar to the first one.

The whole process thus provides a means of transforming a ladder network of the type considered into an equivalent network of the same type, with a modified output impedance. As a special case, the output impedance is unchanged if the product of all the ratios is equal to unity. With the exception of this possible requirement, the arbitrary choice of the transformer ratios is only restricted by physical realizability of the new network elements. In spite of that restriction, a great diversity of equivalent networks can be obtained by this method.

Since the separation between negative and positive values for the new elements is found by equating them either to zero or infinity (on the limit of physical realizability), a suitable choice of the ratios should reduce a number of elements to short or open circuits. This number is generally equal to the number of independent ratios. Which elements can be suppressed by this method is a point to be discussed later.

Formulas will now be considered for the values of the new elements (denoted by primes) under the assumption that all the ratios are known, and without taking into account physical realizability. The network being composed of $n\pi$ sections, there will be n impedance transformations, with respective ratios $k_1, k_2 \ldots k_n$ simultaneously applied. The values of the new elements are obtained from parallel combinations of arms appearing in Fig. 1*B*, and are thus more adequately expressed in terms of admittances. Denoting the admittance of the pth arm by Y_p,

$$Y_p = C_p j\omega + 1/j\omega L_p = C_p j\omega + S_p/j\omega,$$

where S_p is the inverse inductance $1/L_p$, and introducing the notation

$$\left.\begin{array}{l} q_0 = 1 \\ q_1 = k_1 \\ q_2 = k_1 k_2 \\ q_3 = k_1 k_2 k_3 \\ \cdot\ \cdot\ \cdot\ \cdot \\ q_n = k_1 k_2 k_3 \ldots k_n\,, \end{array}\right\} \quad (1)$$

the new admittances are given by the following set of formulas:

$$Y'_{2p} = q_{p-1}\, q_p Y_{2p}\,, \qquad (2)$$

where $p = 1, 2 \ldots n$;

$$Y'_1 = Y_1 + Y_2 - q_1 Y_2; \qquad (3)$$

$$Y'_{2p+1} = q_p[q_p(Y_{2p} + Y_{2p+1} + Y_{2p+2}) - q_{p-1}Y_{2p} - q_{p+1}Y_{2p+2}], \qquad (4)$$

where $p = 1, 2 \ldots n-1$; and

$$Y'_{2n+1} = q_n[q_n(Y_{2n} + Y_{2n+1}) - q_{n-1}Y_{2n}]. \qquad (5)$$

Formula (2) corresponds to series arms, and (4) to shunt arms, the two extreme shunt arms having special expressions, (3) and (5). Each formula can be separated into two similar ones for capacitances and inverse inductances; for instance, (2) gives

$$S'_{2p} = q_{p-1} 1_p S_{2p},\ C'_{2p} = q_{p-1}\, q_p C_{2p}.$$

Consider now the possible suppression of some inductances or capacitances in the modified

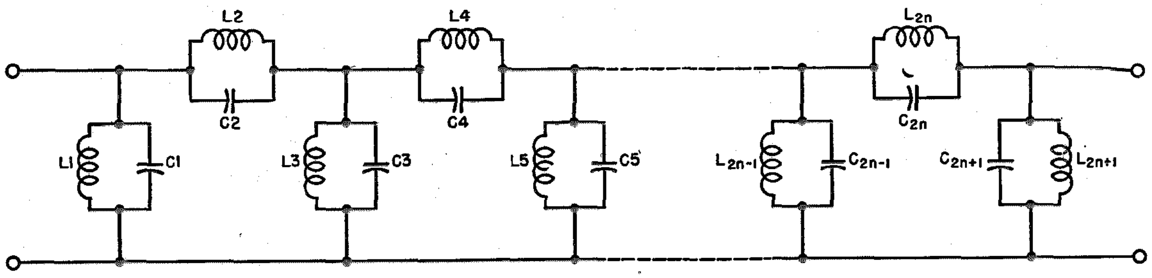

Fig. 4—Ladder network of antiresonant circuits.

network. As general results are difficult to obtain, some simple special cases will be analysed and a qualitative extension of their properties to similar cases will be discussed. We are primarily interested in suppressing elements and not in changing the output impedance, so we will suppose in the following analysis that the over-all transformation ratio q_n is equal to unity.

4. *Transformation of a Double π Network*

The first simple case is a two-section network, represented by the left-hand side of Fig. 4 with $L_5\,C_5$ as the last arm. If the output impedance is unchanged, q_2 must be equal to 1, and (2) to (5) become

$$\left.\begin{aligned} &S_1'=S_1+(1-q_1)S_2\\ &S_2'=q_1S_2\\ &S_3'=q_1(q_1-1)(S_2+S_4)+q_1^2S_3\\ &S_4'=q_1S_4\\ &S_5'=S_5+(1-q_1)S_4\\ &C_1'=C_1+(1-q_1)C_2\\ &C_2'=q_1C_2\\ &C_3'=q_1(q_1-1)(C_2+C_4)+q_1^2C_3\\ &C_4'=q_1C_4\\ &C_5'=C_5+(1-q_1)C_4. \end{aligned}\right\}\quad(6)$$

The central shunt inductance L_3' will be reduced to an open circuit if $S_3'=0$, i.e., if q_1 is determined by

$$q_1=\frac{S_2+S_4}{S_2+S_3+S_4}. \qquad (7)$$

Since this value is less than unity, all the other new elements, except possibly C_3', will be positive. Physical realizability of C_3' requires that

$$q_1\geqq\frac{C_2+C_4}{C_2+C_3+C_4}. \qquad (8)$$

Eliminating q_1 from (7) and (8), this condition becomes

$$\frac{S_3}{C_3}\leqq\frac{S_2+S_4}{C_2+C_4},$$

or

$$L_3\,C_3\geqq(C_2+C_4)\frac{L_2\,L_4}{L_2+L_4}, \qquad (9)$$

and signifies that the antiresonant frequency of the initial central shunt arm $L_3\,C_3$ must be lower than the antiresonant frequency of both series arms $L_2\,C_2$ and $L_4\,C_4$, in parallel. In the limiting case, where equality holds in (9), both S_3' and C_3' vanish. If (9) does not hold, a similar analysis shows that the capacitor C_3' can be suppressed instead of the inductance.

A single π section composed of antiresonant circuits is a classical form of band-pass filter. The frequency of antiresonance of the series arm produces an attenuation peak which may be located at either side of the transmitted band, but the formulas for computing the elements are different in each case. When two π sections are combined, they form a structure of the type considered in this paragraph. The possibility of suppressing an inductance and a capacitance depends on the respective location of the attenuation peaks and the cut-off frequencies.

If both peaks are above the higher cut-off frequency, (9) is satisfied as a consequence of the first theorem of paragraph 2, and the central inductance may be suppressed.

When both peaks are separated by the transmitted band, it can be shown that the special case of equality in (4) is always satisfied, whatever the exact location of the peaks in their respective regions may be. This results directly from the formulas for the elementary sections. The final network in this case is represented in Fig. 5*B*, and is a well-known structure; the method of impedance transformation permits more easily the derivation of the rather intricate formulas for this structure and throws new light on the known fact that it can only be realized when the attenuation peaks lie in two opposite regions.

Final results are summarized in Fig. 5. In the figure, *A*, *B*, and *C* represent the structures obtained in the three cases distinguished above; corresponding attenuation characteristics are sketched in *D*, *E*, and *F*; *G*, *H*, and *I* represent the equivalent inverse T networks with series-resonant arms deduced from the first networks by the duality principle. It should be noted that, as inductors are usually more expensive than capacitors, the series type is to be preferred in the case of *F* and the shunt type in the case of *D* if only attenuation requirements are to be considered.

5. *Transformation of an Iterative Structure*

A second particular case where a complete analysis is possible is the transformation of a midshunt-terminated iterative structure, i.e., a band-pass filter composed of n identical π sections.

Then

$$\left.\begin{aligned} Y_2=Y_4=Y_6=\dots=Y_{2n}=Y_a,\\ 2Y_1=Y_3=Y_5=\dots=Y_{2n-1}=2Y_{2n+1}=Y_b.\end{aligned}\right\}\quad(10)$$

By imposing the condition $q_n=1$, there are $n-1$ ratios $q_1, q_2 \dots q_n$ to be determined by the suppression of some of the elements. In suppressing the $n-1$ intermediate shunt inductances in the modified structure, $n-1$ linear equations are obtained for the ratios:

$$q_p(2+S_b/S_a)-q_{p-1}-q_{p+1}=0, \quad (11)$$

where $(p=1,2\dots n-1)$.

This set of equations will be considered as a linear difference equation for q_p with respect to the index p, the boundary conditions being $q_0=q_n=1$. By writing

$$\left.\begin{aligned} 1+S_b/2S_a=\cosh\varphi\ ,\\ \text{or}\qquad \sinh\varphi/2=\sqrt{S_b/2S_a}\ ,\end{aligned}\right\}\quad(12)$$

the solution works out in compact form:

$$q_p=\frac{\cosh\left(\frac{n}{2}-p\right)\varphi}{\cosh n\varphi/2}. \quad (13)$$

As all values of q_p lie between 0 and 1, the positive value of only the intermediate shunt capacitances $C'_{2p+1}(p=1,2\dots n-1)$ need be checked. One easily obtains the formula

$$C'_{2p+1}=q_p^2C_b(1-f_b^2/f_a^2), \quad (14)$$

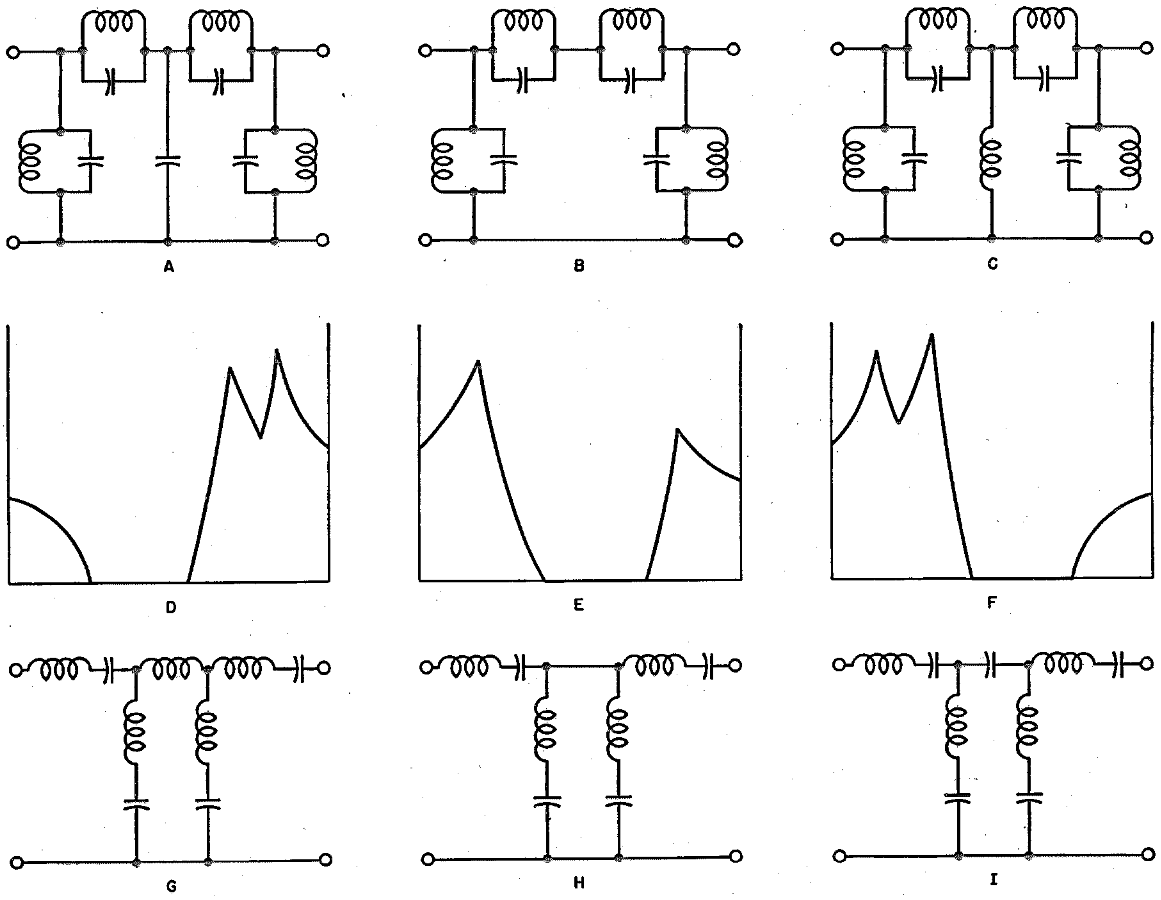

Fig. 5—*A*, *B*, and *C* are single π networks composed of antiresonant elements. *D*, *E*, and *F* are graphs of attenuation versus frequency, and *G*, *H*, and *I* are the equivalent T networks. The equivalent π and T networks and their attenuation characteristics are arranged vertically.

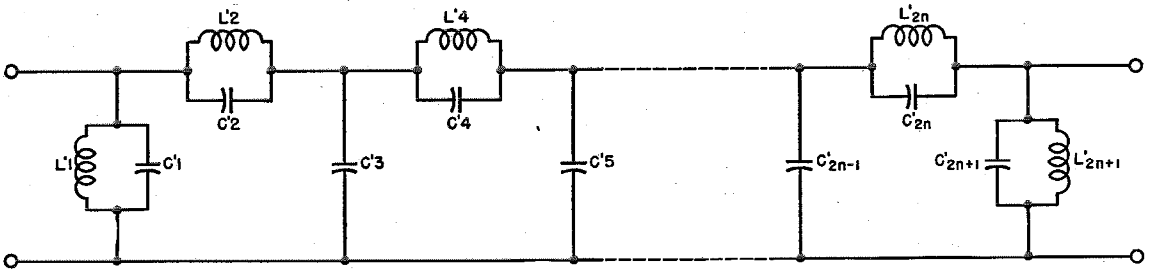

Fig. 6—Structure similar to a low-pass filter derived by suppressing the intermediate shunt inductances.

where f_a and f_b are the respective antiresonant frequencies of the original series and shunt arms. As a consequence of (14), the intermediate shunt inductances can be suppressed if $f_a > f_b$, i.e., if the frequency of infinite attenuation, common to all sections, lies above the transmitted band. The iterative filter is then transformed into a structure represented in Fig. 6, and explicit formulas may be obtained in terms of the usual design parameters. Notations and formulas are listed below:

$f_1, f_2 =$ cut-off frequencies $(f_1 < f_2)$.

$f_\infty =$ frequency of infinite attenuation $(f_\infty > f_2)$.

$$m = \sqrt{\frac{f_\infty^2 - f_2^2}{f_\infty^2 - f_1^2}} = \tanh \varphi/2.$$

$n =$ number of sections.

$R =$ nominal filter impedance.

$$L_0 = \frac{R}{2\pi(f_2 - f_1)}.$$

$$C_0 = \frac{f_2 - f_1}{2\pi R f_1^2}.$$

$$x = f_1/f_\infty.$$

$$q_p = \frac{\cosh (n/2 - p)\varphi}{\cosh n\varphi/2}, \ (p = 0,1,2 \ldots . n).$$

$$L'_{2p} = \frac{2mL_0}{q_p q_{p-1}}, \ (p = 1,2 \ldots . n).$$

$$C'_{2p} = q_p \, q_{p-1} \, x^2 C_0 \cosh^2 \varphi/2, \ (p = 1,2 \ldots . n).$$

$$C_1' = C'_{2n+1} = \frac{C_0}{2}[(1 - q_1)x^2/m + \sinh \varphi].$$

$$L_1' = L'_{2n+1} = \frac{2mL_0}{\cosh \varphi - q_1}.$$

$$C'_{2p+1} = q_p^2 (1 - x^2) C_0 \sinh \varphi, \ (p = 1,2 \ldots . n - 1).$$

6. *Conclusions*

From the preceding analysis, it should be expected that in any composite band-pass filter for which the various infinite-attenuation frequencies are about the same and are all located above the transmitted range, all intermediate shunt inductances can be suppressed by simultaneous impedance transformations with an overall ratio equal to 1. This is also roughly suggested by the structure of Fig. 6, which is very similar to that of a low-pass filter. Although all attempts to prove this statement for an arbitrary alloca-

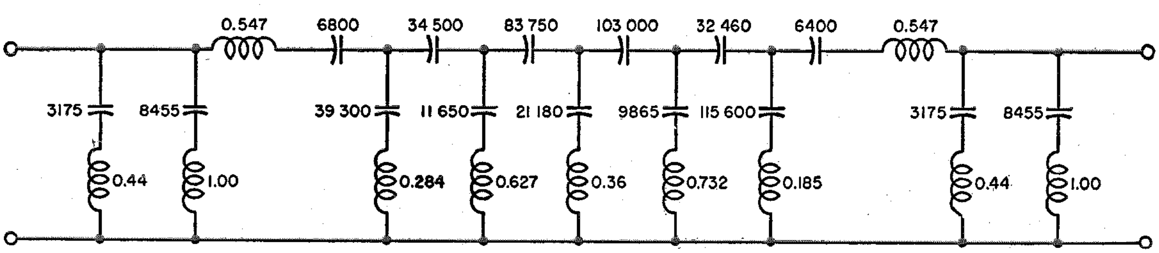

Fig. 7—Band-pass filter of 600 ohms impedance passing frequencies from 60.6 to 108 kilocycles. Inductances are in millihenries and capacitances are in micromicrofarads.

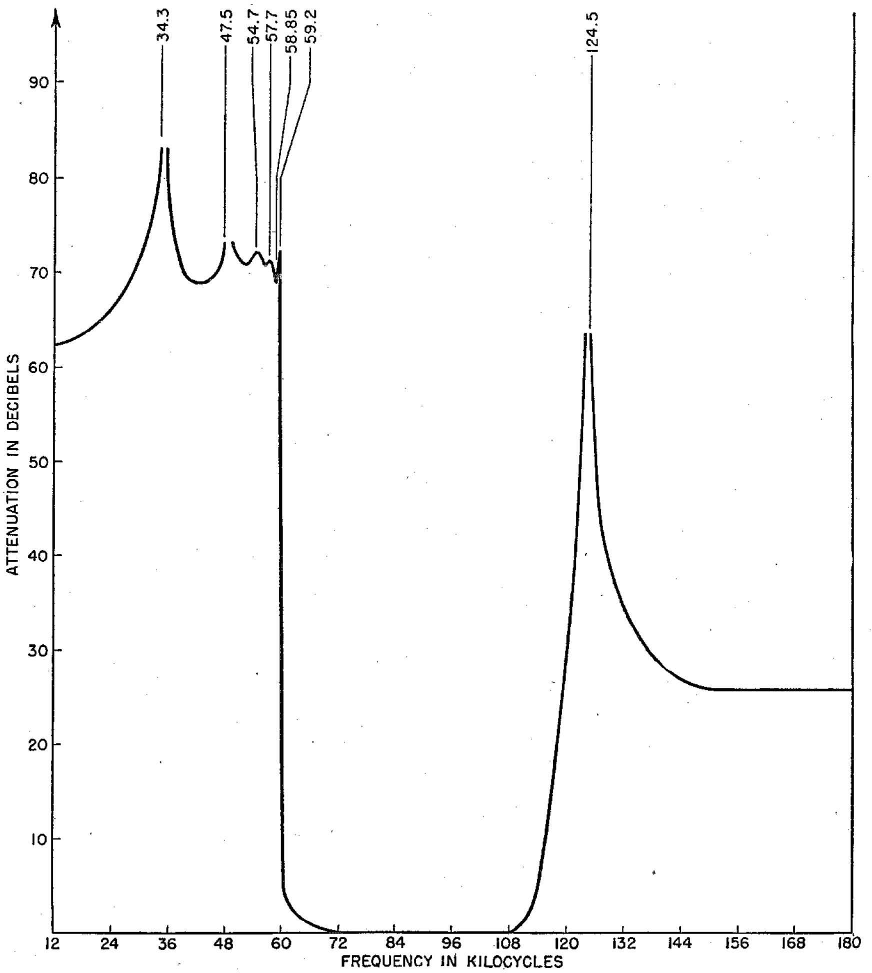

Fig. 8—Measured insertion loss of filter shown in Fig. 7.

tion of attenuation peaks have been unsuccessful, its correctness was practically confirmed by the author's own experience in designing such filters.

By the duality principle and frequency transformation, the suppression of intermediate series inductances should be possible in a composite band-pass filter of the inverse type when all the frequencies of infinite attenuation lie below the transmitted band.

An illustrative example of this last case will be discussed briefly. A band filter (60.6–108 kilocycles per second, 600 ohms) having high attenuation (60 decibels in the range 12–59.4 kilocycles) for the lower side, and only a moderate attenuation (about 20 decibels) above 120 kilocycles, was required. This filter is represented in Fig. 7, and the measured insertion loss in Fig. 8. Attenuation peaks at 124.5 and 54.7 kilocycles, and *m*-derived terminal impedances, are produced by two terminal half-sections of the classical type. The central part of the filter, producing attenuation peaks in the lower range only, has been successfully transformed by the use of the present theory.

Bikini Observations and Their Significance*

By HARADEN PRATT and ARTHUR VAN DYCK

Official United States Scientific Observers

THE OBSERVERS of Operation Crossroads were accommodated on ships of the general communications type. Designed to carry large military, naval, and air staffs in amphibious operations, they provided ideal quarters for this purpose. The observers were of several kinds: scientific men from the United States and the United Nations, United States congressional representatives, officers of the United States War and Navy Departments, and representatives of the press. In spite of the size and complexity of Operation Crossroads, it was executed in most efficient fashion throughout, and the excellent handling of observers was but one example of the general effectiveness of the organization, which was under the able command of Vice Admiral Blandy.

The *U. S. S. Panamint*, to which we were assigned, arrived at Bikini lagoon on the morning of June 29th and immediately steamed to an anchorage a short distance from Bikini Island by very slowly moving through the entire target array of some 72 ships. The sight was impressive in all respects. A mighty fleet was anchored row upon row on the azure waters of an immense tropic lagoon fringed with palm tree islands, waiting for the awesome test scheduled only two days away. Battleships, cruisers, carriers, destroyers, submarines, transports, and ships of other smaller types—even to a concrete drydock—were variously grouped around the target's bull's-eye to which the eye continually reverted, namely, the majestic battleship *Nevada*, conspicuous in bright orange paint and white turret tops and guns, the better to guide the bombardier on the fateful day.

It was obvious that these ships had been prepared for this special event since they were disposed in a pattern intended to reveal maximum information on damage at all distances. Their decks were fitted with all kinds of equipment and materials of war to be subjected to the explosion. Each ship had graduated scales painted on bow and stern so that settlement could be noted from time to time through observation from aircraft. Even the islands bore evidence of the vast preparation that preceded our visit, as several steel towers to accommodate cameras and other instruments were easily visible.

That afternoon we visited the *Nevada*, the Japanese battleship *Nagato*, and the carrier *Independence*, where we saw the large number of test specimens mounted on their decks. These included samples of clothing, food, armor plate, airport fuel trucks, medical supplies, airplanes, and hundreds of other items of military weapons and supplies. The *Nagato*, commissioned about 1921, was of particular interest with her war wounds of the two direct hits from aerial bombs, including the skip bomb that went right through the ship via Admiral Yamamoto's quarters.

Long studies of Bikini July weather indicated that perhaps as many as twenty days could elapse before a suitable one for the drop would arrive. Besides adequate visibility it was required that the wind blow in the same direction from sea level to 20 000 feet so that radioactive products of the atomic fission would move away from all observers and not endanger them in the event of rain. However, on June 30th such weather was predicted for July 1, and the fleet of some 140 attending vessels steamed out of the lagoon maneuvering during the night to be at their assigned locations for the big event at 8:30 the next morning.

We on the *Panamint* saw the blast from the bridge deck through very dark special goggles at a point about twenty miles from the *Nevada*. Not being able at this distance to see any of the target fleet, many of us were not looking directly at the correct spot and so missed the initial point of flash. By the time eyes had moved over, the burst had already become a disk somewhat larger than the sun and considerably brighter, a conclusion made possible by being able to glance at the sun several times before the ship's public-address system announced "bomb away."

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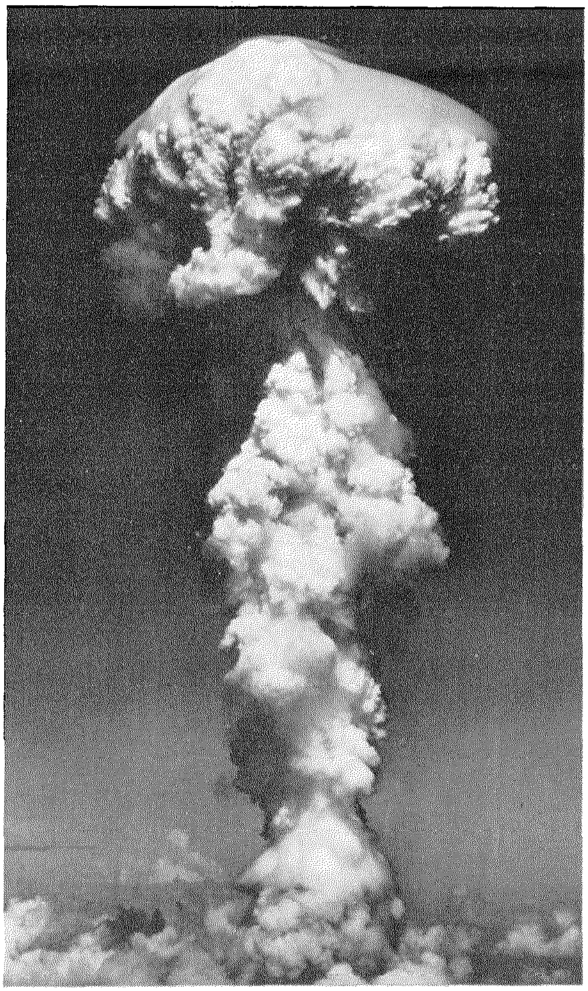
* Reprinted in part from *Proceedings of the I.R.E. and Waves and Electrons*, v. 34, pp. 930–933; December, 1946. Presented at a joint meeting of the Institute of Radio Engineers, American Institute of Electrical Engineers, and Radio Club of America in New York, New York, on November 6, 1946. All photographs in this article courtesy of Joint Army-Navy Task Force One.



The first Bikini bomb, detonated above the water, sent a billowing cloud of smoke and flame almost forty thousand feet into the air.

The disk of intense light was immediately blotted out by the instantaneously formed luminous dome or hemisphere of incandescent gases which rested on the water, covering much of the target area. Quick loss of luminosity occurred, and with bare eyes we viewed the majestic column of atomic cloud with its mushroom top rise and shoot several thousand feet skyward in a matter of seconds. We estimated that this structure rose to a total of some 38 000 feet, displaying from the start interesting shades of pinkish colors against a fleecy white. The yellowish aspect of previous atomic explosions, ascribed to dirt and debris, was, of course, absent at Bikini.

By eleven that morning, we had moved up to the reef and examined the target fleet through binoculars while seeing at the same time yellow-colored drone boats darting about picking up water samples to test for radioactivity. These boats were remotely controlled from a distant destroyer with air units observing and directing. Test results were favorable for certain parts of the lagoon and our ships took anchorage there soon after lunch. Here we were able to survey the fleet clearly and note the many wrecked superstructures. The outstanding spectacles were the Japanese cruiser *Sakawa* with a list, down at the stern, and a completely wrecked top, and the carrier *Independence* with bad fires which culminated toward evening in spectacular explosions leaving the ship a shambles. The *Sakawa* turned turtle and sank the next morning.

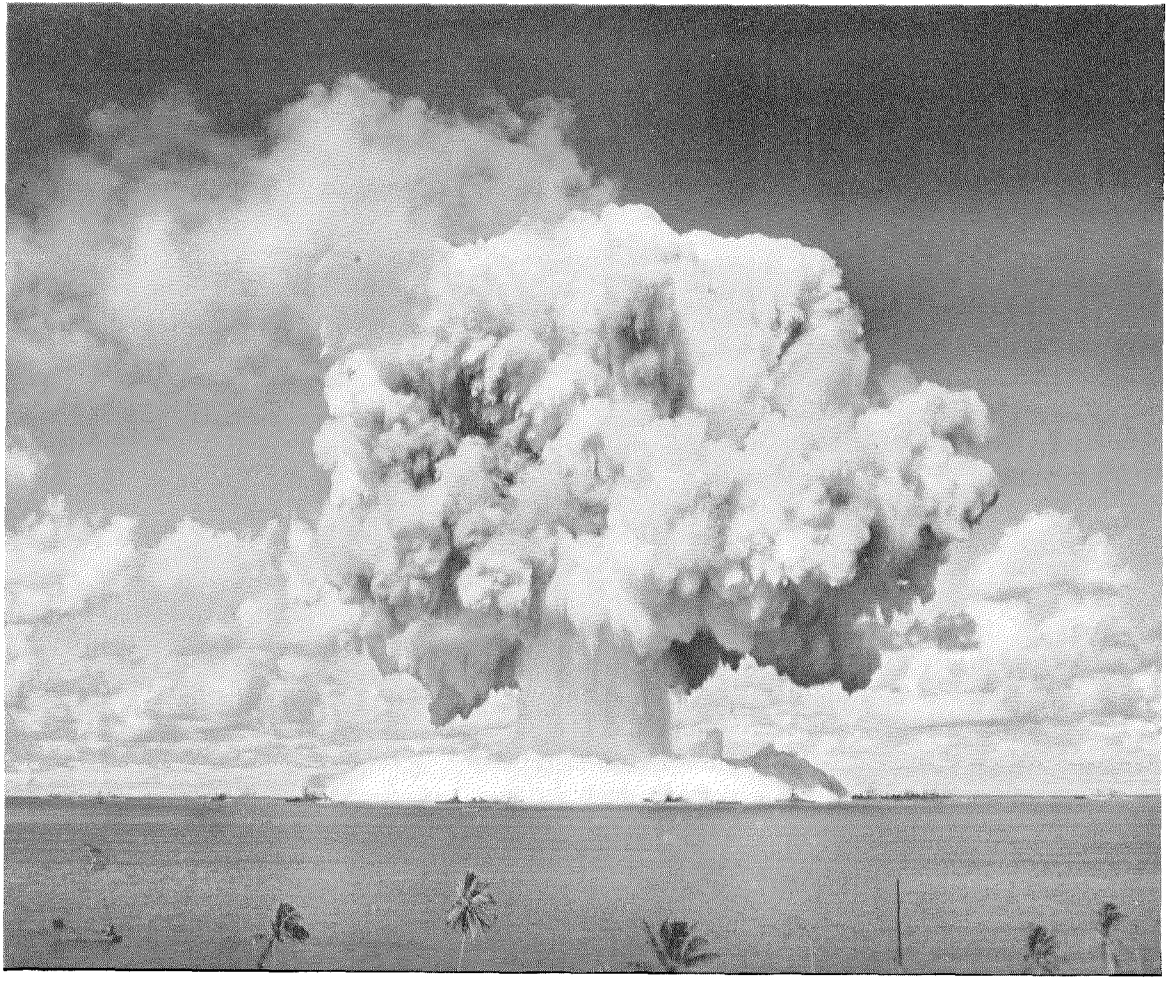
Interesting and spectacular as all these events were, the full realization of the enormous significance of what had taken place unfolded rather slowly during succeeding days as we visited and examined ship after ship. Lessened radioactivity enabled ships to be boarded 48 hours after the burst. Within three-quarter mile radius, exposed wood was scorched black, crates and boxes were burned, and the *Nevada's* after deck, hit by the blast at an angle of about 25 degrees, was crushed down and blackened. Her funnels were pushed into her superstructure and the airplane crane on her stern bent double. It should be explained that after striking bull's-eyes on many practice runs, the bombing plane had the hard luck, on the real drop, of missing the *Nevada* by some hundreds of feet, a sore disappointment to the Army Air Forces.

Conditions were the same on the *Arkansas*, and worse on the *Pensacola* which was within a half-mile radius. All these vessels' decks and superstructures were a mass of wreckage, with bent bulkheads, twisted railings, smashed doors, stacks down, antenna gear deformed or broken—not to speak of peeling paint from the heat wave and the damaged or burned-out specimens placed on their decks. The blast wave even penetrated below, wrecking furnishings and doing other damage in spots here and there. Many vital items were seriously damaged, such as bulkhead-mounted motor-control cabinets, the switches and other parts of which were broken loose and completely inoperative. Broken castings in quantity taught that naval ships of the future must avoid the use of metal fabricated in brittle forms.

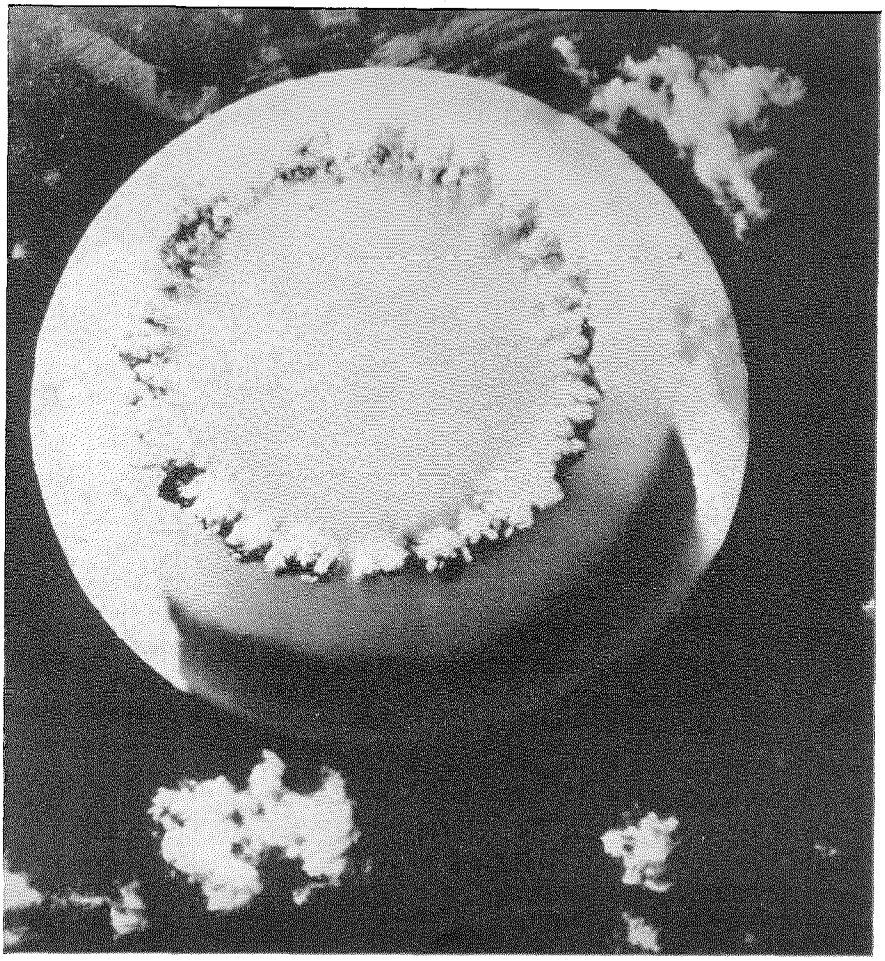
It was unfortunate that our public was misled during the weeks preceding the tests by statements appearing in the press forecasting dire results and the possible unleashing of forces of nature such as earthquakes, tidal waves, and volcanic eruptions. It must be remembered that the effects of explosions diminish very rapidly with distance, and while the atomic holocaust sank and wrecked vessels up to one-half mile, damage to ships beyond one mile was relatively light. Heat, blast, and wave action at Bikini Island three miles away left almost no visible traces. Even direct blast and heat damage from huge volcanic explosions such as Krakatoa and Katmai extended over only small areas. Certain excited spectators, therefore, had no logical basis for reporting disappointment because Bikini trees were not uprooted or because the blast at twenty miles did not blow them off their feet.

It must be remembered that, while five ships were sunk in the air test and about nine sunk plus two beached in the underwater test, conclusions as to the power of atomic bombs should not be based on the number of ships sunk. Obviously, if the whole fleet had been closely bunched most would have been sunk, whereas if widely dispersed not more than one such casualty would have occurred.

However, all these arresting phenomena, significant in their seriousness as they are, represent only the effect of heat and explosive action arising from the concentration of stupendous power at



The huge column of radioactive water lifted by the atomic bomb was just starting to fall as this picture was taken by an automatically operated camera located on the atoll.



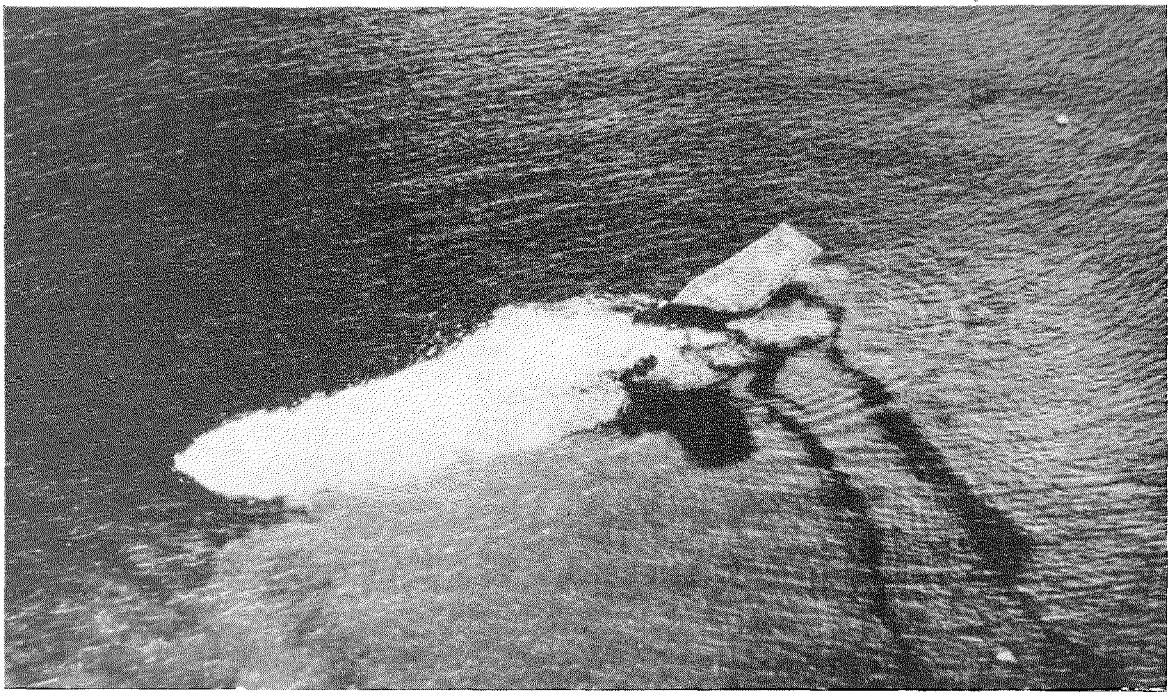
An airplane flying almost directly above the underwater explosion took this picture of the radioactive water and gasses which surged upward in a column a half mile in diameter.

a single point. The effects of radioactive radiations, primarily neutrons and gamma rays, constitute the new feature which justifies the appellation of "Poison Bomb" as conveying the real meaning of this colossal development. Even though radiation diminished so fast after the first test as to enable ships to be safely visited a few hours later, it is doubtful whether but a few could have survived had the ships been manned when the bomb was dropped, even though the bulk of the crews might have lived long enough to render ships operative and resist post-explosion attack.

Any doubts as to the sweeping nature of the poison effect which may have existed were removed after the second atom bomb was exploded below the surface of Bikini lagoon. Millions of tons of sea water, hurled skyward over one mile in a column almost a half mile in diameter, were heavily contaminated with the fission products estimated as equivalent to hundreds of tons of radium. In the first test these products distributed in the atmosphere and were dispersed by the winds. This death-laden water in the second test fell directly on all ships in the lagoon and engulfed some in waves 70 to 100 feet high. This resulted in radioactive products being washed down ventilators, pipes, funnels, and scuppers, in saturation of all topside hamper, and in penetration to hidden places such as circulating systems, pumps, and evaporators.

Even ships not in the target fleet became dangerously contaminated by entering affected lagoon areas afterwards. Many of the surviving vessels are still uninhabitable and others present vexatious problems of reconditioning because of gamma rays coming from materials lodged in the scale and incrustations inside sea-water piping.

Much has been said as to whether the tests were necessary and their cost justified, even though the cost was less than the value of a single modern battleship. Our considered opinion is that the tests were indispensable. Aside from arguments that scientists might fairly well predict results, and apart from the tremendous value of the precise technical information acquired, the mere holding of the operation under controlled conditions with many observers from all walks of life was of incalculable value to all mankind because the impact of the atom bomb reaches



Seven and a half hours after the underwater explosion, the gallant old aircraft carrier *Saratoga* slipped beneath the radioactive waters of Bikini lagoon.

human beings everywhere. And there is no substitute for actual results indisputably to drive home the facts.

The significance of the Bikini tests is clear and powerful. Nevertheless, it has been the universal experience of the United States observers, on return from Bikini, that people with whom they have discussed the matter have been uncertain as to the significance of the tests and of the atom bomb generally. We have even found many people who are unwilling to talk or think about the subject, saying that it is just too horrible to contemplate. We would like to convince such people, and all people, that the atom bomb not only is horrible, but that it is so terrible that something must be done about it. That 'something is not to hide our heads in the sand—it is that we must insure ourselves against its use. And that means we must somehow prevent all war in the future.

The facts are very clear, and the best presentation may be merely to list them in simple language.

An atom bomb of the present type, exploding in the air, destroys everything within about one-quarter mile, does very heavy damage to one-half mile, and heavy damage to one mile. Beyond one mile, the degree of damage depends upon the character of structures. Windows and light structures will be shattered at several miles.

An atom bomb of the present type kills practically all the human beings within one-quarter mile, a very high percentage of those within one-half mile, and a great many of those within one mile or more.

The present-type bomb, bursting in the air over New York City, would blow out every window within one or two miles and would knock off most of the roof structures and brick and stone facings of buildings, particularly skyscrapers. Casualties from glass and falling debris would be high. Fires from short circuits, broken gas mains, and other causes would be numbered in hundreds.

The present-type bomb, bursting under the water in New York City, would destroy subways, and would render uninhabitable for months an area of at least ten square miles. Each seaport city of the country would be similarly exposed.

The atom bomb is not the only new weapon of vastly greater destructiveness. The guided missile, like the German V-1 and V-2, is another.

The power and destructiveness of weapons has been increasing rapidly for the last 100 years. World War I saw the first wide use of high explosives.

World War II achieved vast destruction. Most people of the United States do not realize this. The people of London, Coventry, Rotterdam, Warsaw, Stalingrad, Berlin, Tokio, and Pearl Harbor *do* realize it.

World War II dislocated civilization, and almost completely wrecked it through destruction of so much of the economic structure of the world.

A third World War will be vastly more destructive of both economic structures and of human beings. Since World War II was almost enough to destroy civilization, a vastly worse War III is certain to do so.

There is no defense against the atom bomb or against the guided missile, or, of course, against a combination of the two. Defense has never been perfect against any weapon. Against the atom bomb, unless the defense is perfect, it is no defense. Not one German V-2 missile was shot down of the many that approached London. If two or three of them had had atom bombs in them, London would have become an empty shell.

We have now reached that advanced state of civilization wherein we have made it possible for a few uncontrolled members of our society to destroy or to subjugate the rest of us, before we can do anything about it. The fact is that material development has reached a dangerously high level. We have been settling arguments by force from the beginning of man on earth, but usually the side of moral right has been able to marshal enough might to prevail sooner or later. Now we have a new situation, and there is little protection left in material things.

All this has been said before. Indeed, it has been said so many times that the thought has become familiar and has lost its true meaning to many people. This complacency is dangerous. These facts mean that a revolution in human affairs has occurred. Hereafter civilization must struggle not to advance, but actually to survive. These are not mere words, they are elementary truths. All this should be obvious to everyone, and particularly to those men charged with the responsibility of government. It is evident that the only safety is in means which will make it impossible for any nation to attack another.

The only possible safe policy for the future must be one which rests on the law and the conscience of man. Nations must be controlled, as we now control states, counties, cities, and individuals.

We must have world law and order. That is the simple significance of nuclear fission—and of Bikini.

Landing Aircraft with Ground Radar

By J. S. ENGEL

Federal Telephone and Radio Corporation, Newark, New Jersey

GUIDANCE in the landing of aircraft under conditions of poor visibility and without special airborne equipment of any kind is the function of Radio Set AN/MPN-1C, a mobile equipment. A search radar system picks up the airplane at distances to 30 miles, and at 10 miles control is shifted to a precision radar system and a team of three operators. Standard air-ground radiotelephone communication is used to inform the pilot constantly of his position until, at 50 feet above the runway, the pilot is ordered to complete the landing without further aid.

• • •

Ground radar for informing a pilot of the position of his aircraft in making an approach for a blind landing was first suggested in 1941 by Dr. Luis W. Alvarez of the Radiation Laboratory at Massachusetts Institute of Technology. His proposal was stimulated by the needs of the air forces. The radio range and marker-beacon systems then in use required additional airborne receiving and indicating equipment as well as special training of pilots—handicaps which the air forces could ill afford. By 1943, the new ground-controlled-approach equipment, developed under the direction of Dr. Alvarez, and later to be manufactured in the U.S.A. by Federal Telephone and Radio Corporation and others, was operating sucessfully in England, guiding bombers, back from raids over Germany, to safe landings through zero visibility.

As embodied in Radio Set AN/MPN-1C, radar is used to track incoming aircraft, which may carry only standard airborne two-way communication equipment to receive all necessary information. The set has given dependable service under most types of instrument landing conditions. Its mobility is an added feature, permitting rapid shifting from one runway to another and from one airfield to another, as

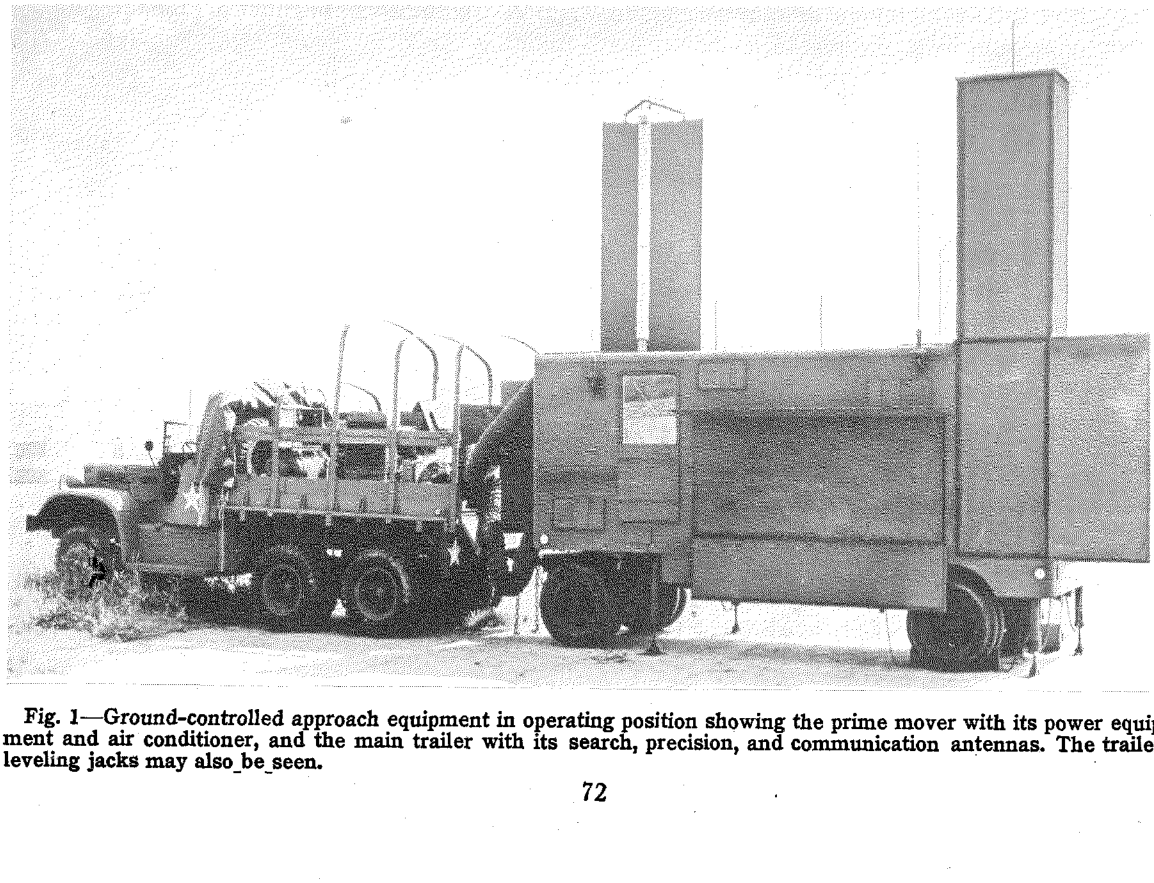


Fig. 1—Ground-controlled approach equipment in operating position showing the prime mover with its power equipment and air conditioner, and the main trailer with its search, precision, and communication antennas. The trailer-leveling jacks may also be seen.

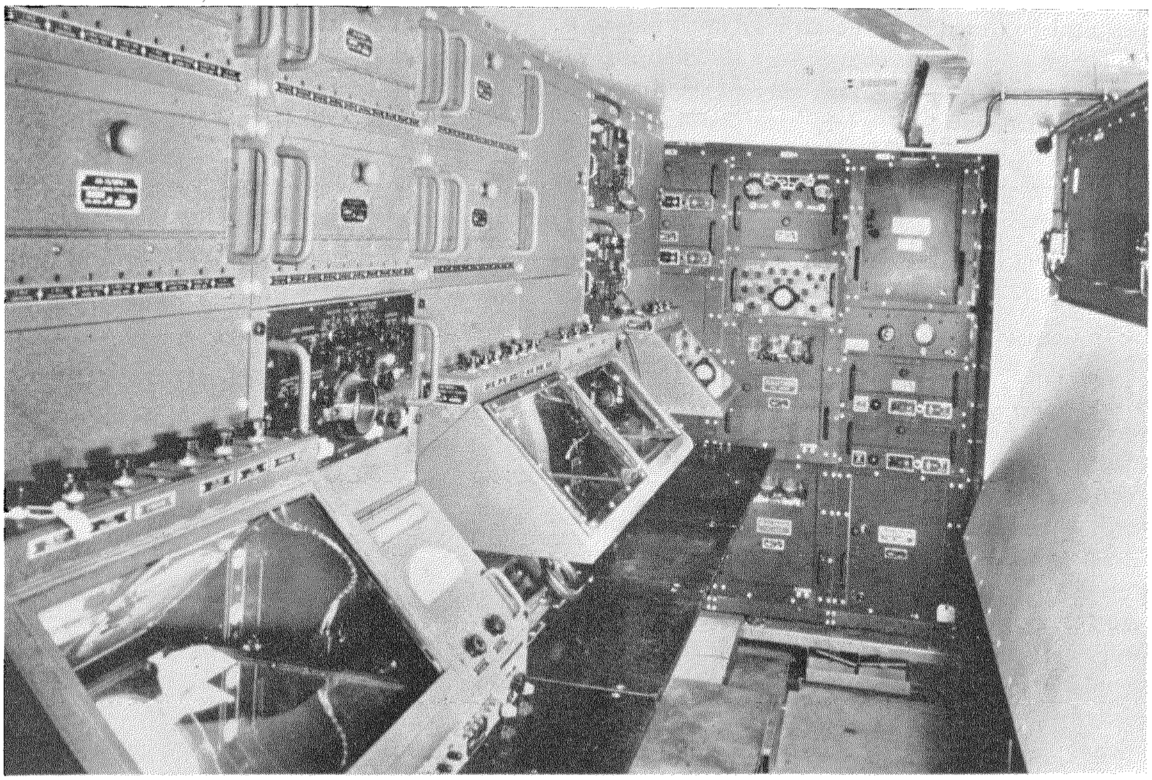


Fig. 2—Interior of the trailer. The various indicators, synchroscopes, radar transmitters, receivers, transmission lines, and communications equipment are much in evidence.

conditions require. Fig. 1 shows the equipment set up for operation.

1. *Operation*

The ground equipment comprises a search radar, a precision radar, and standard two-way radiotelephone air-ground facilities installed in a single trailer which is located near the windward end of a landing runway. The precise location of the trailer is determined with respect to the beam-scan coverage of the precision-radar azimuth antenna. The search radar obtains information for controlling air traffic within a 30-mile radius at altitudes up to 4000 feet. The precision radar supplies data to direct an aircraft along the proper glide path for a safe approach. Each radar, operating in the microwave range, has its own transmitting, modulating, receiving, and indicating systems. The search-radar antenna scans 360 degrees in azimuth. The two antennas of the precision radar scan sectors in azimuth and elevation, respectively. A synchronizer unit triggers the modulators and indicators of both radars. Further to insure synchronization, the pulse modulator is common to both radars. The communications equipment is used to relay information to the aircraft being landed. Fig. 2 is an interior view of the trailer.

1.1 LANDING PROCEDURE

An aircraft entering the traffic-control area of the airport establishes radio contact with the control tower and requests landing instructions. The operator in the tower transfers control of the aircraft to the traffic director in the trailer. Viewing the plan-position indicator, which is part of the search radar, the traffic director first identifies the aircraft and then, by radio, instructs the pilot as to correct heading and elevation.

A common method of establishing identity is to have the aircraft fly along a leg of a nearby radio range so as to come in directly over the range station, which the pilot identifies by the cone of silence above it. The pilot informs the traffic director as soon as the aircraft enters the cone of silence. After verifying this position on the plan-position indicator, the traffic director gives the pilot headings which enable him to fly an orbital course about the airfield until clearance for an approach is signaled by the operator in the control tower. In this way, several aircraft may be managed while awaiting landing instructions. As all aircraft flying in the vicinity of the airfield are visible as radar targets, the traffic director is enabled to give the instructions needed to avoid collisions.

1.1.1 Search Radar

The presentation on the plan-position indicator is in the form of a polar map having the radar at the center and showing range and azimuth. Distances from the radar to the various targets are measured along the radii of the sweep; the azimuth indication is synchronized with the beam direction of the search-radar antenna. The plan-position indicator may be operated on any one of three ranges: 7.5, 15, or 30 miles. It is not viewed directly, but through a semisilvered mirror mounted in a horizontal position. The indicator tube is mounted at a 45-degree angle above the mirror, and a map (which shows the proper approach to the field) is mounted at a 45-degree angle below the mirror. As seen through the mirror, the map appears to be superimposed on the face of the indicator tube.

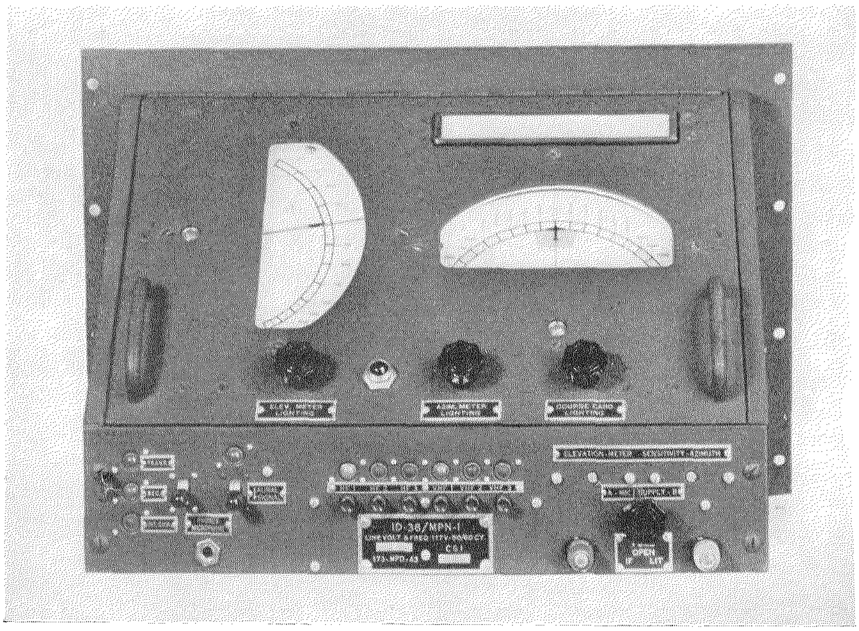


Fig. 3—A view of the approach-controller's indicator showing the error meters and communication-channel-selection controls.

When the tower signals that the field has been cleared for an approach, the radio communications channel controlled by the traffic director is taken over by the plane selector, who works from another plan-position indicator in the trailer. The plane selector then gives one of the waiting aircraft instructions for starting an approach. The pilot is given headings and altitudes until the plane is within range of the precision radar. The starting position is approximately at a 1000-foot altitude, 8 to 9 miles from the field and in line with the end of the runway. The plane selector then directs the pilot to head toward the runway and, over the intercommunication system in the trailer, informs the operators of the precision radar that the aircraft is ready to complete the approach.

1.1.2 Precision Radar

The precision operators comprise a team of three men: an azimuth tracker, an elevation tracker, and an approach controller. By operating the handwheel cursors and pedals which control antenna positions, the azimuth tracker follows the aircraft first on the 10-mile indicator, and then on the 2-mile indicator. During the last part of the approach he places the "fast-slow scan" switch in the "fast" position, which gives him greater tracking accuracy. To facilitate resolution of the aircraft from other targets at this close range, the azimuth tracker also operates the receiver gain control, enabling him to regulate the amplitude of the received echo. The elevation tracker performs similar operations at the 10- and 2-mile elevation indicators, except that he is relieved of the control of the scan rate.

Like the plan-position indicator of the search radar, the indicators of the precision radar are viewed through semisilvered mirrors. The azimuth and elevation indicators employ four 7-inch cathode-ray tubes. The two azimuth indicators show asimuth with respect to range as determined by the azimuth-scanning antenna, and the two elevation indicators show elevation with respect to range as determined by the elevation-scanning antenna. This type of presentation is known as expanded partial plan-position indication. On this indicator the angular coverage is expanded, and the trace origin is located near the edge, instead of at the center, of the face of each tube thus utilizing a greater area of each tube face.

The intensity-modulated picture which appears on each tube face reveals the radar-calibrating reflectors which are placed at appropriate positions along the runway, the obstructions and terrain along the glide path, as well as the approaching aircraft. The azimuth and elevation trackers see each radar picture with a superimposed illuminated cursor and a V-notch follower. The cursor, on which a line is engraved, is mounted so that it can be moved parallel to the range marks on the radar map. The V-notch follower is mounted so as to pivot about the trace origin. When the aircraft signal comes into view each tracker first bears on it with the V-notch follower by working the antenna-control pedals. The azimuth tracker's pedals control the elevation antenna, while the elevation tracker's pedals control the azimuth antenna. Each tracker then turns the handwheel which adjusts the cursor so that the engraved line is centered on the received signal. If the aircraft changes course, he readjusts the pedals and handwheel to keep the cursor and follower "on target."

When both trackers are definitely "on target" they operate the "data good" switches which illuminate the error meters on the approach indicator. (See Fig. 3.) The error meters are operated by the cursors on the indicators of the azimuth and elevation trackers, are calibrated in feet, and indicate deviations of the aircraft from the desired path. This information is interpreted by the approach controller.

After receiving "good data," the approach controller is ready to take over the control of the aircraft and "talk down" the pilot. Using the intercommunication system, the approach controller informs the plane selector accordingly, and takes over control of the radio communication channel to give the pilot preliminary instructions for completing the approach. These include landing drill (wheels and flaps), correct air speed, and proper rate of letdown. The approach controller continues to relay correct headings and rate of descent to the pilot so that he remains in the proper path. Towards the end of the approach the pilot is directed not to use his radio to acknowledge instructions. When the aircraft is about 50 feet above the runway, the approach controller tells the pilot how many feet he is from the center line of the runway and orders him to complete the approach visually.

Thus from the instant that the pilot establishes standard radio contact with the ground-controlled-approach equipment, his navigation is accomplished for him from the ground, except for the last few feet of his course when he can see his way to a visual landing. Fig. 4 illustrates the manner in which the equipment performs its functions of guiding aircraft to a safe landing under conditions of zero visibility.

2. *Equipment*

To insure dependable operation, all electronic components of the search and precision radars, except for indicators and antennas, are duplicated as two complete channels. Either channel may be operated while the other is on stand-by or off. Normally, one channel is operated at a time. Should the operating channel fail, the stand-by channel is placed in service, and the defective channel is secured for repairs.

Each channel of the search radar includes a modulator, transmitter-converter, and receiver. Both channels employ the same antenna and indicators.

Each channel of the precision radar includes a synchronizer which triggers the search radar, a modulator, a transmitter-converter, a receiver, two sweep amplifiers, two channel switch units, two angle-coupling units, and a set of power supplies. The antenna assembly and indicators are common to both channels of this radar.

The communications system comprises three very-high-frequency radio transmitter-receiver equipments (SCR-522-A), each transmitter of which may be operated on any one of four pre-tuned channels between 100 and 156 megacycles with 8 to 10 watts of power and amplitude modulation, and three high-frequency radio transmitter-receiver equipments (SCR-274-N transmitters and BC-342 receivers). Each of these transmitters may be operated on any one of three pretuned channels between 2 and 9 megacycles with 15 to 25 watts of power and amplitude modulation.

2.1 SEARCH RADAR

The 10-centimeter search radar is fired by the synchronizer unit, which is part of the precision radar. A blocking-oscillator circuit in the synchronizer generates a positive trigger voltage at a repetition rate of 2000 cycles per second. This trigger voltage is fed to a line-controlled blocking oscillator in the modulator, which generates a 0.5-microsecond positive pulse. The pulse is applied to the grids of two pulse-modulator tubes connected in parallel to a capacitor which has been charged to 15 kilovolts through a resistive network from the high-voltage power supply. When the modulator tubes conduct, the capacitor is discharged, so that a negative 15-kilovolt pulse is applied to the magnetron which in turn generates a 10-centimeter radio-frequency pulse of 0.5-microsecond duration. After each pulse, the capacitor recharges. In this way the magnetron transmits a radio-frequency pulse of 85 to 100 kilowatts of peak power (85 to 100 watts average power) 2000 times each second.

Energy from the magnetron is coupled through a duplexing circuit and a transmission line to the antenna. The function of the duplexer is to switch the single antenna from the transmitting circuit to the receiving circuit. Only a small part of the total transmitted energy which strikes the target is reflected to the antenna and coupled through the duplexer to the receiving circuit.

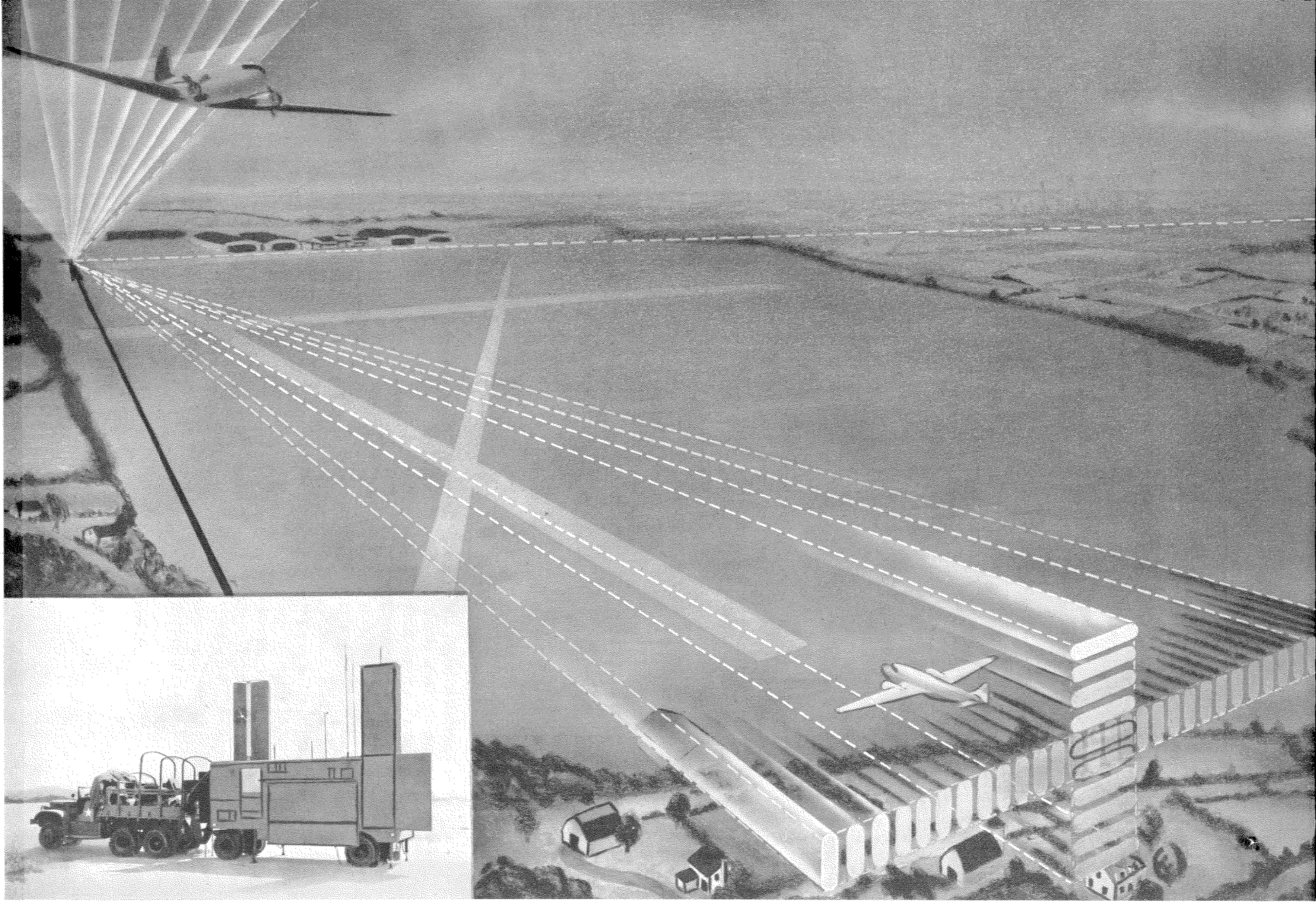
In the receiver, a McNally klystron generates a continuous-wave signal which differs from the magnetron frequency by 30 megacycles. This signal is heterodyned with the received signal in a crystal mixer. The resultant 30-megacycle beat frequency is coupled to a preamplifier located near the mixer, and then to the signal intermediate-frequency amplifier. The preamplifier is also coupled to the automatic-frequency-control circuits which maintain a constant difference of 30 megacycles between the klystron and magnetron frequencies regardless of frequency drift in the magnetron or klystron. After detection, the video-frequency signal is applied to the intensity grid of the cathode-ray tube of each search indicator.

The search central of each search-indicator assembly generates the sweep voltages, range marks, and intensification gate applied to the indicator tube. These voltages are initiated by a negative trigger voltage from the synchronizer. The voltages from a sine-wave voltage divider, which is mechanically coupled to the search-antenna drive, are mixed with the sweep voltages and the resultant voltage applied to the indicators. A single radial sweep is thus initiated each time the transmitter fires. A direct voltage is applied to the sine-wave voltage divider, and as the antenna scans, the divider output consists of two voltages, which are functions of the search-antenna position and are 90 degrees out of phase with each other. The angular data are thus transmitted electrically to the search indicators.

2.1.1 *Antennas*

The antenna is mounted on a motor-driven assembly installed in the roof of the trailer. This drive rotates the antenna at 30 revolutions per minute, so that it scans 360 degrees every 2 seconds. Energy is coupled to the antenna through a rotating coaxial joint. The antenna consists of a vertical array of 10-centimeter dipoles coupled to a $1\frac{1}{2}$- by 3-inch wave guide, and mounted along the focal line of an 8- by 4-foot reflector of parabolic cross section. The dipoles are spaced and energized so as to produce a vertical radiation pattern having sharp cutoff below the horizon and a cosecant-squared characteristic above the horizon. When the antenna is adjusted so that the nose of the vertical beam just clears the horizon, an area within a range of 30 miles and up to an altitude of 4000 feet can be scanned. The horizontal cross section of the beam generated by the reflector is symmetrical and less than 7 degrees wide at its half-power points.

Fig. 4—On the opposite page is a view of an airport and environs showing the relative position of the ground-controlled-approach equipment and the sky coverage provided by the search and precision systems.



The merit of the cosecant-squared pattern is that there is no gap in coverage at higher altitudes and close range. All targets, regardless of range, are thus scanned more uniformly than they would be by a pencil beam. Another advantage of the pattern is the sharp cutoff at the horizontal which reduces ground clutter and permanent echoes.

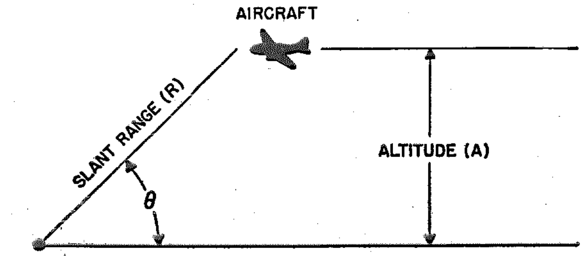


Fig. 5—The slant range of an aircraft flying at a constant altitude is equal to A cosec θ.

The cosecant-squared characteristic of the antenna distinguishes it from other types of antennas generally used with air-search radars. Most such equipments are designed to give early warning, and therefore use antennas whose sharp pencil-shaped beam patterns are directed near the horizon. Good echoes are thus received from targets at low altitude and considerable range, but not from targets at high altitude and close range. To overcome the danger of losing aircraft orbiting close to the airfield at an altitude of 4000 to 5000 feet, the cosecant-squared pattern has been made a characteristic of the present antenna.

As shown in Fig. 5, when the slant range R of the aircraft is large, the angle of elevation θ is very nearly zero. However, if the altitude A of the aircraft approaching the radar remains constant, angle θ increases to an appreciable value. The slant range R is equal to $A/\sin\theta$ or, since the cosecant of an angle is equal to the reciprocal of its sine, to A cosec θ. As θ becomes large, i.e., greater than 1 degree, the transmitted power is propagated through free space, and the received power becomes directly proportional to the square of the gain G of the antenna, but inversely proportional to the fourth power of R. If a constant C is used to represent the other factors involved, the echo power P_r becomes

$$P_r = C\frac{G^2}{R^4},$$

therefore

$$G = R^2\sqrt{\frac{P_r}{C}}.$$

As $R = A \operatorname{cosec}\theta$,

$$G = A^2 \operatorname{cosec}^2\theta\sqrt{\frac{R}{C}}.$$

As it has been assumed that the aircraft is approaching at a constant altitude and that an echo of uniform amplitude is required, it is convenient to consider $A^2\sqrt{P_r/C}$ as a constant K. Therefore, $G = K \operatorname{cosec}^2\theta$. Thus to receive an echo of uniform amplitude from an aircraft approaching at a constant altitude, irrespective of its range, it is necessary to use an antenna whose gain in the vertical plane is proportional to the square of the cosecant of the angle of elevation. This function is obtained in the search-radar antenna by adjusting the spacing between the dipoles and the amount of power coupled to each dipole.

2.2 Precision Radar

The precision radar operates in the 3-centimeter band at a repetition rate of 2000 cycles per second. The 15- to 20-kilowatt, 0.5-microsecond radio-frequency pulse transmitted by the magnetron, with a duty cycle of 1/1000, through the duplexer to the antenna system is generated in the same way as in the search radar. The precision radar, however, employs two antennas; one comprising an azimuth-scanning array and the other an elevation-scanning array.

Energy from the transmitter-converter is fed alternately to each antenna for approximately 0.125 second. This is accomplished by the radio-frequency switch, which consists of a waveguide Tee whose branch arms are alternately shorted. Shorting is accomplished by semicircular metallic blades inserted alternately between two pairs of choke flanges which are separated by an air gap. The blades are mechanically coupled by a shaft. Radio-frequency energy is transmitted across each gap without loss when a blade is removed. The length from the center of the Tee to each pair of choke joints is adjusted so that when a blade is inserted all the energy is fed to the open arm. Associated with this switch is a commutator which unblanks the indicator connected to the energized antenna.

2.2.1 Antennas

The precision antennas must scan rapidly because of the high speeds at which modern aircraft land. Antennas of the early equipments were designed for mechanical scanning but their mass was found to be too great. In Radio Set AN/MPN-1C, electrical scanning at two speeds of 1 and 4 scans per second is used. The dipole arrays of the precision antennas are coupled to wave-guide sections so that the axis of each array is along the line of focus of a semiparabolic reflector. This produces a very narrow beam in the plane parallel to the axis of the array (as determined by the length of the antenna). In the axis normal to the array, the beam width is determined by the width of the reflector.

The azimuth antenna, which is mounted in the trailer so that the axis of its array is in the horizontal plane, radiates a horizontally polarized beam whose half-power width is 1 degree in azimuth and 2 degrees in elevation.

The elevation antenna, which is mounted in the trailer so that the axis of its array is in the vertical plane, radiates a vertically polarized beam 0.6 degree wide in elevation and 3.6 degrees wide in azimuth at the half-power points. Fig. 6 shows the installation of the antennas in the trailer.

When electrical energy is propagated through a wave guide at a given frequency, the velocity and wavelength of the energy in the guide are functions of the physical dimensions of the guide. If the frequency is kept constant and the dimensions of the wave guide are changed, the guide wavelength λ_g is changed in direct proportion to the dimensions, as expressed by

$$\lambda_g = \frac{\lambda}{\left\{\dfrac{1 - \lambda^2\left[\left(\dfrac{m}{a}\right)^2 + \left(\dfrac{n}{b}\right)^2\right]}{4}\right\}^{\frac{1}{2}}},$$

where a and b are the depth and width, respectively, of the wave guide, and m and n are subscripts denoting the particular mode of propagation. The scanning technique employed by the precision antennas is based on this formula.

Each precision antenna comprises an array of dipoles spaced at fixed distances and coupled to a length of wave guide whose dimensions can be easily changed to vary the phase of the energy exciting each dipole. The angular position of the beam of each precision antenna is thus shifted with respect to the axis of its array by changing the wave-guide dimension. A cam drive coupled to the radio-frequency switch mechanically increases and decreases the broad dimension of the wave guide. The sector which is scanned by the beam of each precision antenna is therefore a function of the motion of the cam drive. The scanning mechanisms are constructed so that the horizontally polarized beam of the

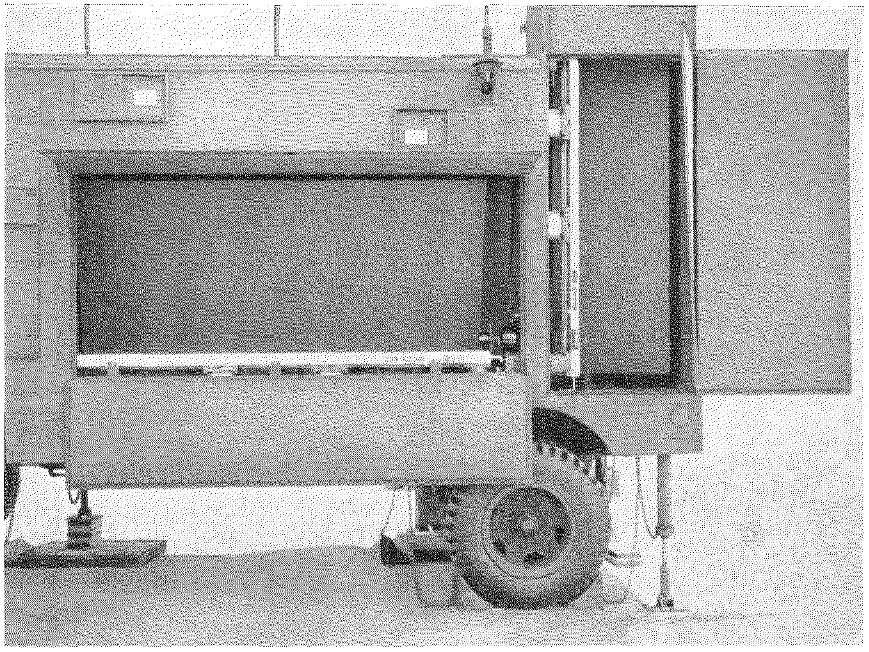


Fig. 6—Precision antennas mounted in the trailer. The azimuth antenna (horizontal) and elevation antenna (vertical) are shown in their compartments.

azimuth antenna and the vertically polarized beam of the elevation antenna sweep alternately. While one beam sweeps the other is unenergized.

During a complete cycle, the scanning sequence is as follows; the azimuth antenna scans from 5 degrees to the right to 15 degrees to the left of the runway; the elevation antenna then scans from 1 degree below to 6 degrees above the horizon; the azimuth antenna scans back from 15 degrees to the left to 5 degrees to the right of the runway; and the elevation antenna completes the cycle by scanning back from 6 degrees above to 1 degree below the horizon.

To obtain reasonable gain from the arrays, the associated reflectors must be quite large. This reduces the beam width of each array along the nonscanning axis. To prevent loss of signal when the aircraft deviates appreciably from course (that is when it is flying in a weak part of the antenna beam), the antenna-directing pedal system is provided to rotate the elevation scanner in azimuth and azimuth scanner in elevation.

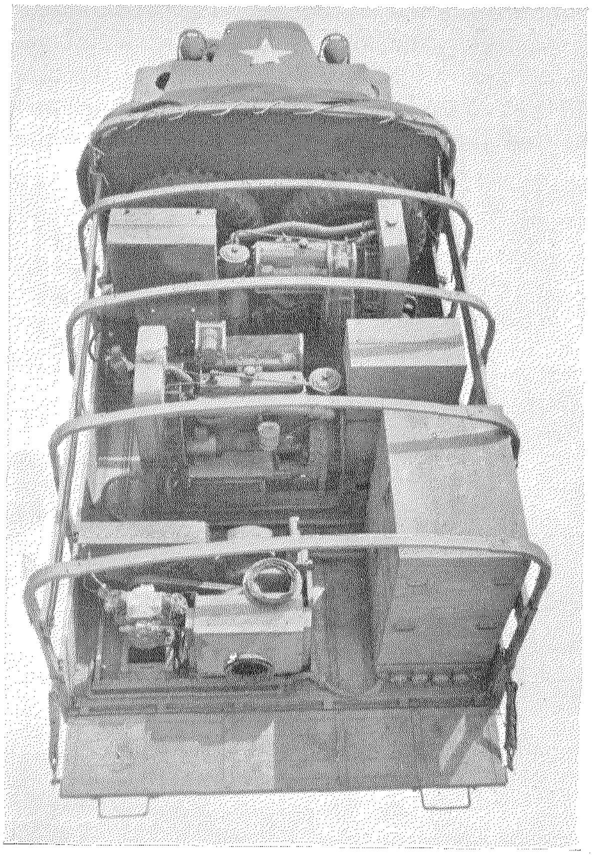


Fig. 7—A view of the prime mover showing the installation of the engine-generator sets and the air-conditioning equipment.

As in the search radar, the received signals from targets in the area scanned by the precision antennas are fed from the antennas to the receiver through the duplexer. As the precision radar operates in the 3-centimeter band, a Shepperd-Pierce klystron serves as a local oscillator instead of the McNally tube used with the 10-centimeter search radar. The receiver of the precision radar is identical with that of the search radar. The receiver is provided with a sensitivity-time-control circuit, which is triggered by the synchronizer, and varies the sensitivity of the intermediate-frequency as the fourth power of the time following each pulse (corresponding to the fourth power of range). The effect is uniform signal amplitude from targets of the same size and aspect, regardless of the target range. The sensitivity-time-control circuit is not used in the search receiver. The receiver video-frequency output is fed to the synchronizer, where an amplifier and cathode-follower couple the signals to the precision indicators.

Horizontal and vertical electromagnetic deflection coils are mounted on the neck of each indicator tube. A single sweep-generator-amplifier is used to drive both cathode-ray tubes of each indicator. Trigger voltages are supplied by the synchronizer, and antenna-position data are supplied to each indicator by an angle-coupling unit which delivers a direct voltage, the amplitude of which is a function of the position of the antenna beam. This voltage modulates the amplitude of the sweep currents in the deflection coils, and thus transmits the instantaneous position of the antenna beam to the indicator map.

The elevation and azimuth antennas are each equipped with an angle-coupling unit. This arrangement is duplicated for the stand-by channel. The angle-coupling units compensate for the nonlinear scanning mechanisms of the antennas. Each unit consists of a capacitive dividing network, mechanically coupled to its respective scanning mechanism and electrically fed from a 1-megacycle oscillator.

As the precision antennas operate alternately and the same receiver feeds both azimuth and elevation indicators, the indicator which is not receiving data must be blanked until the associated antenna is used. This is accomplished by a phototube blanking commutator which is mechanically coupled to the antenna-scan drive. Thus when the elevation antenna is energized, the elevation indicator is unblanked, and when the energy is switched to the azimuth antenna, the azimuth indicator is unblanked. The rapid scan rate makes it difficult to use a mechanical switch for unblanking the indicators. Instead, a phototube light-beam switch is used. Two blanker switch blades coupled to the radio-frequency switch in the antenna drive are used alternately to interrupt a light beam and thus unblank the appropriate indicator.

The gain of the elevation antenna differs from that of the azimuth antenna, and since the same transmitter-receiver is used with both antennas, it is necessary to control the system gain for each part of the cycle independently. A gain control on each indicator is connected to the receiver for this purpose by the phototube switching unit when an indicator is unblanked. The 2500-foot and 2-mile range marks applied to the indicators are generated in the synchronizer.

The error voltages which are applied to the meters of the approach indicator are obtained from nonlinear voltage dividers which are mechanically coupled to the tracking cursors on the precision indicators. In aligning the system for operation, adjustments are made so that the error meter will read zero when the cursor is set by the handwheel on the proper glide path.

3. Mechanical and Power Supplies

All electronic equipment is mounted in and on a trailer with a total weight of about 21 000 pounds. In setting up for operation, hydraulic jacks are used for leveling. The trailer is 237 inches long, 98 inches wide, and 125 inches high. It is hauled by a prime mover which carries two engine-driven generators and air-conditioning equipment. (See Fig. 7.) The loaded tractor weighs approximately 24 000 pounds and is 268 inches long, 96 inches wide, and 128 inches high. On the road, the coupled vehicles will turn in a radius of less than 37 feet.

The engine-driven generators supply 117/220 volts at 60 cycles. Two 9-kilovolt-ampere generators operating at power factors between 0.73 and 0.85 deliver a maximum of 13.5 kilowatts.

Frequency-Modulated Broadcast Transmitters for 88–108 Megacycles

By LEONARD EVERETT

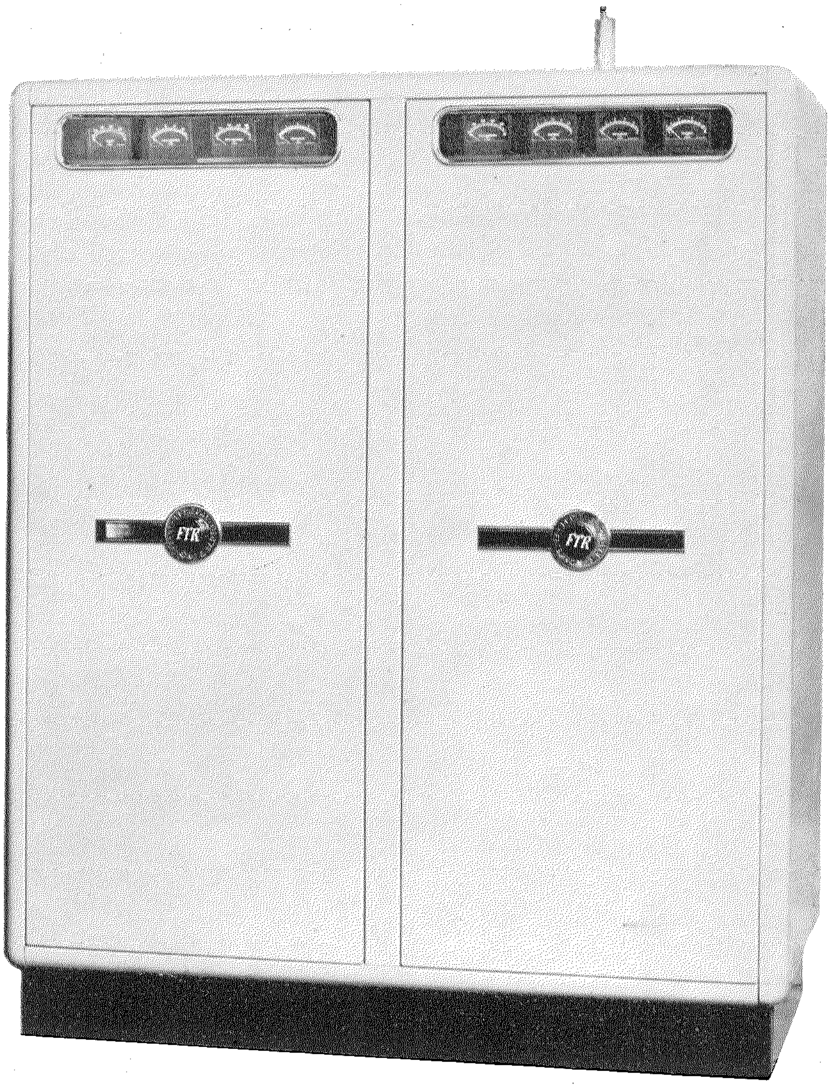
Federal Telephone and Radio Corporation, Newark, New Jersey

DESIGNATION of the frequency band from 88 to 108 megacycles per second for frequency-modulated broadcasting in the U.S.A., required the design of completely new transmitting equipment. Transmitters of 1, 3, 10, 20, and 50 kilowatts output are based on the "Frequematic" modulator which provides for wide frequency swings for modulation but prevents drifting from the assigned center frequency. Use of multiple square-loop antennas permits radiation in the vertical plane to be concentrated at the horizon while maintaining a circular azimuthal pattern.

• • •

Frequency-modulated broadcast transmitters have been designed for output powers of 1, 3, 10, 20, and 50 kilowatts. A still wider range of effective radiation results from the use of multiple square-loop antennas[1] having useful power gains up to 12 times that of a simple dipole antenna. The power gain at the horizon over the field produced by a dipole is approximately equal to the number of square loops stacked vertically. The use of 12 loops, giving a power gain of 12, is practicable. Thus, it is possible to select the most economical transmitter and antenna combination to meet conditions for any particular installation.

The performance of a frequency-modulated transmitter is dependent to a very large extent on the type of modulator and center-frequency-stabilization system employed. Use of the all-electronic Frequematic modulator and center-frequency-stabilization system permits these



3-kilowatt frequency-modulated transmitter, front view, doors closed.

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[1] R. F. Lewis, "Square Loops for Frequency-Modulated Broadcasting at 88–108 megacycles," *Electrical Communication*, v. 23, pp. 415–425; December, 1946.

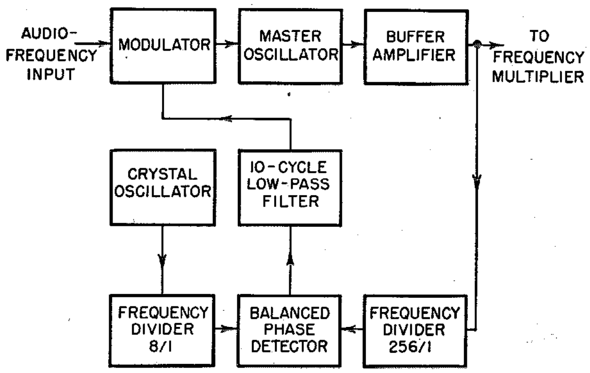


Fig. 1—Block diagram of Frequematic modulator and center-frequency-stabilization system.

transmitters to operate well within the standards set by the Federal Communications Commission for frequency-modulated broadcasting in the U.S.A.

1. *Electrical Characteristics*

The center-frequency-stabilization system associated with the Frequematic modulator literally locks the center frequency in synchronism with the frequency of a temperature-controlled quartz crystal. This piezoelectric oscillator is maintained at a frequency within 0.001 percent of its required value, resulting in the transmitter center frequency being held to within ±1000 cycles.

In accordance with standards of the Radio Manufacturers Association, the normal program input level to the transmitter is approximately 0 volume units for full modulation. A 75-microsecond pre-emphasis circuit is provided at the audio-frequency input. Use of this circuit is optional and it can be readily disconnected if pre-emphasis is provided ahead of a limiting amplifier, or if it is desired to check the response-frequency characteristics of the transmitter. The response-frequency characteristics of the transmitter follow the standard 75-microsecond pre-emphasis curve to within 1 decibel between 50 and 15 000 cycles. The distortion, including harmonics up to 30 kilocycles, is less than 1.5 percent of the root-mean-square values for fundamental frequencies between 50 and 15 000 cycles; and below 1.0 percent for fundamental frequencies between 100 and 7500 cycles.

The frequency- and amplitude-modulated noise present in the transmitted signal, measured under normal operating conditions, are at least 65 and 50 decibels, respectively, below full carrier modulation.

2. *Modulator*

Fig. 1. shows a block diagram of the Frequematic modulator and center-frequency-stabilization system. The audio-frequency input signal is applied to a modulator tube which, by means of the well-known Miller effect, produces a capacitive variation across the tank of a master oscillator, these variations being proportional to the applied audio-frequency signal. Fig. 2 is a schematic illustration of this circuit, from which it may be seen that the audio-frequency signal is applied to the grid of the modulator tube. The signal produces a variation in the transconductance of the tube with a resultant variation in its input capacitance. By proper selection of tube type and circuit components, the variation in input capacitance is made proportional to the amplitude of the applied signal. From Fig. 2 it may be seen that the grid of the modulator tube is coupled through a capacitor C_c to the tank circuit of a Hartley oscillator. Capacitor C_{gp},

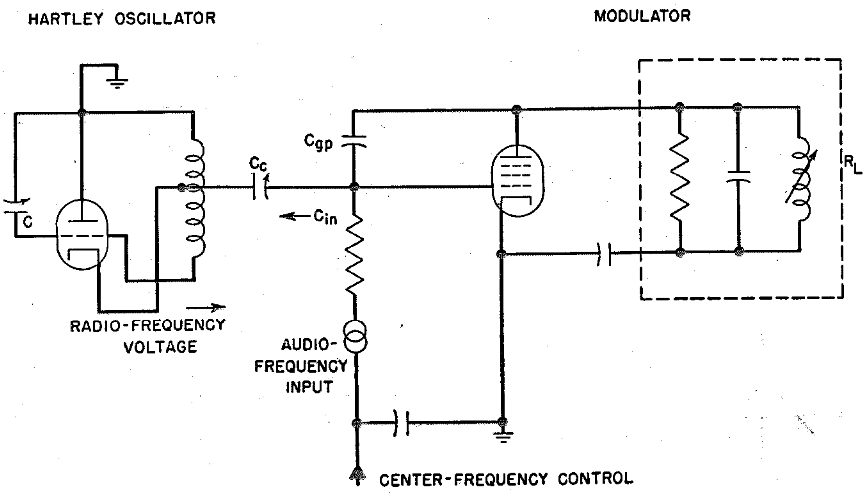


Fig. 2—Basic circuit of modulator.

which is connected between the grid and plate, serves to add to the interelectrode capacitance and to magnify the Miller effect produced in the tube.

To function properly, the modulator plate circuit must be tuned to the resonant frequency of the oscillator, because only under the conditions of resonance will a purely capacitive impedance be reflected onto the grid of the tube. If the plate circuit is detuned, the impedance reflected onto the grid becomes partly resistive, resulting in the introduction of amplitude modulation. For this reason, the modulator plate tank circuit has been given a relatively low Q, making its tuning broad and thus preventing any appreciable detuning effect as a result of the shift in frequency of the oscillator with modulation. Use of the low-Q circuit also makes the tuning noncritical and greatly facilitates adjustment of this stage.

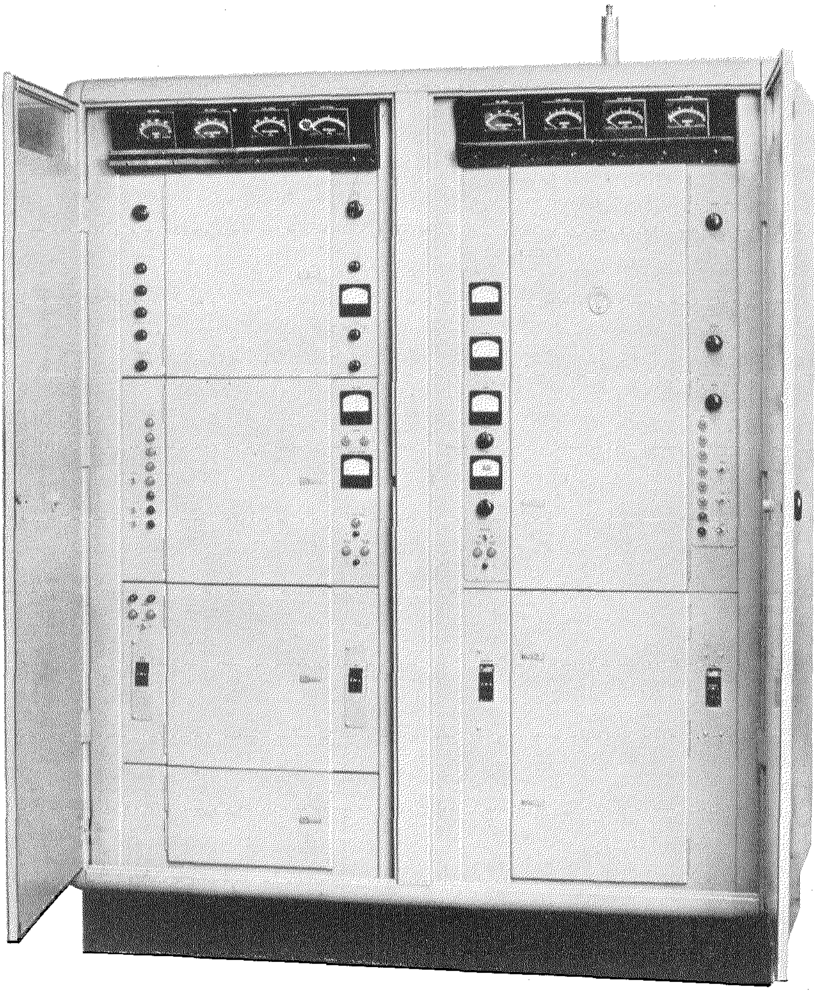
The oscillator stage is of the conventional Hartley type, but with a grounded plate and the cathode tapped across the tank inductance. The tank circuit is temperature compensated to reduce frequency drift to a minimum. A variable powdered-iron core in the tank inductance permits frequency adjustment between 3.4 and 4.7 megacycles. The output of the oscillator is fed through a buffer amplifier stage to the frequency-multiplying stages of the transmitter, where a multiplication of 24 times produces the required transmitter output frequency.

3. Center-Frequency Stabilization

From the block diagram of Fig. 1, it will be seen that a portion of the buffer-stage output is applied to a frequency-dividing network. This consists of a series of multivibrators which reduce the frequency by a factor of 256 and produce an output in the neighborhood of 15 kilocycles. This output is applied to a balanced phase detector.

The output of a crystal-controlled oscillator, after being divided in a similar network by a factor of 8, is also applied to the balanced phase detector. The frequency of the crystal oscillator is such that, when the master oscillator is operating at its proper frequency, the two voltages applied to the phase detector will be of the same frequency. This is accomplished by employing a crystal having 1/768 and a master oscillator of 1/24 of the transmitter output frequency.

Fig. 3 shows the basic circuit of the balanced phase detector. The output of this circuit is ap-



3-kilowatt frequency-modulated transmitter, front view, outer doors open.

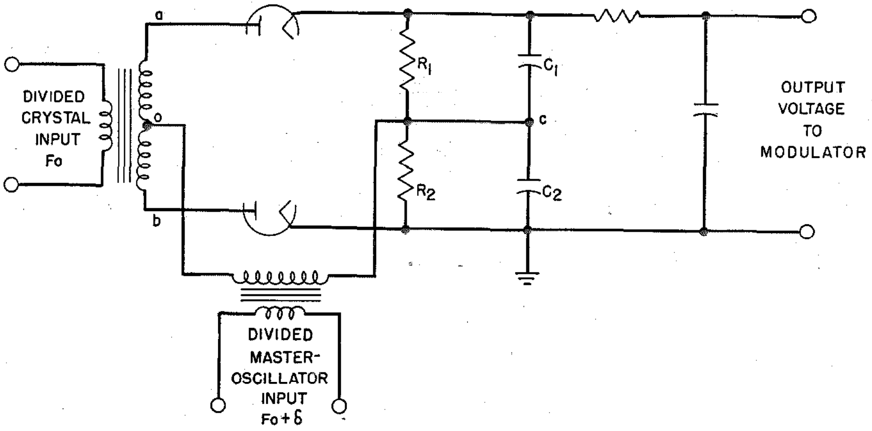

a
o
b
DIVIDED CRYSTAL INPUT Fo
R1
R2
C1
C2
c
OUTPUT VOLTAGE TO MODULATOR
DIVIDED MASTER-OSCILLATOR INPUT Fo+δ

Fig. 3—Basic circuit of balanced phase detector.

plied to the grid of the modulator tube in series with the audio-frequency input signal. Under normal operating conditions, the two voltages applied to the phase detector are identical in frequency and differ only in relative phase. For purposes of explanation, assume that this condition exists. The two voltages will then add vectorially, producing a resultant voltage which is rectified and applied to the modulator tube.

The amplitude and polarity of the voltage applied to the modulator tube will depend on the relative phase of the two voltages as shown by the curve of Fig. 4. From this curve it will be seen that, when the two voltages have a phase relationship of $\pi/2$ radians, or 90 degrees, the resultant output voltage is zero. Other phase relationships will cause a positive or negative correcting voltage to be applied to the grid of the modulator tube. The application of a correcting voltage to the grid of the modulator tube causes a change in the amount of capacitance injected across the tuned circuit of the master oscillator. The correcting voltage acts on the modulator in the same manner as the applied audio-frequency input voltage, except that the correcting voltage, being a direct voltage, tends to change the frequency in one direction only. As in the case of the audio-frequency input voltage, the capacitive change in the oscillator tank is proportional to the amplitude of the correcting voltage.

The arrangement of the circuit is such that the application of a correcting voltage to the modulator will tend to change the tuning of the oscillator in a direction which reduces the amplitude of the correcting voltage. When the master oscillator is operating at its proper frequency, the voltages applied to the phase detector will be of the same frequency, and will have a definite phase relationship. Any attempt of the master oscillator to drift will produce a change in the phase of the voltages applied to the phase detector, and a resultant change in the capacitance injected into the tank circuit of the master oscillator. This corrects the tendency to drift and maintains the master oscillator at its original frequency. As a result of this action, the output voltage of the master oscillator can do no more than change very slightly in phase with respect to the crystal oscillator. Thus the frequency of the master oscillator must remain constant and is effectively locked in synchronism with the frequency of the crystal oscillator.

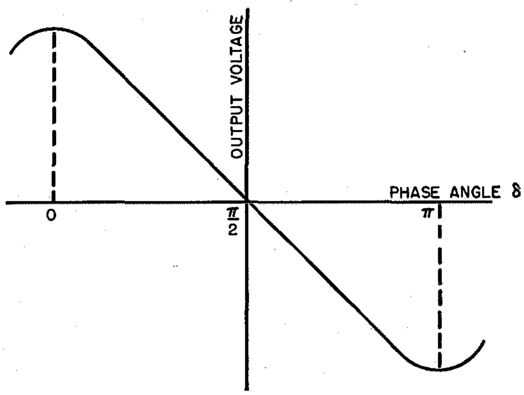
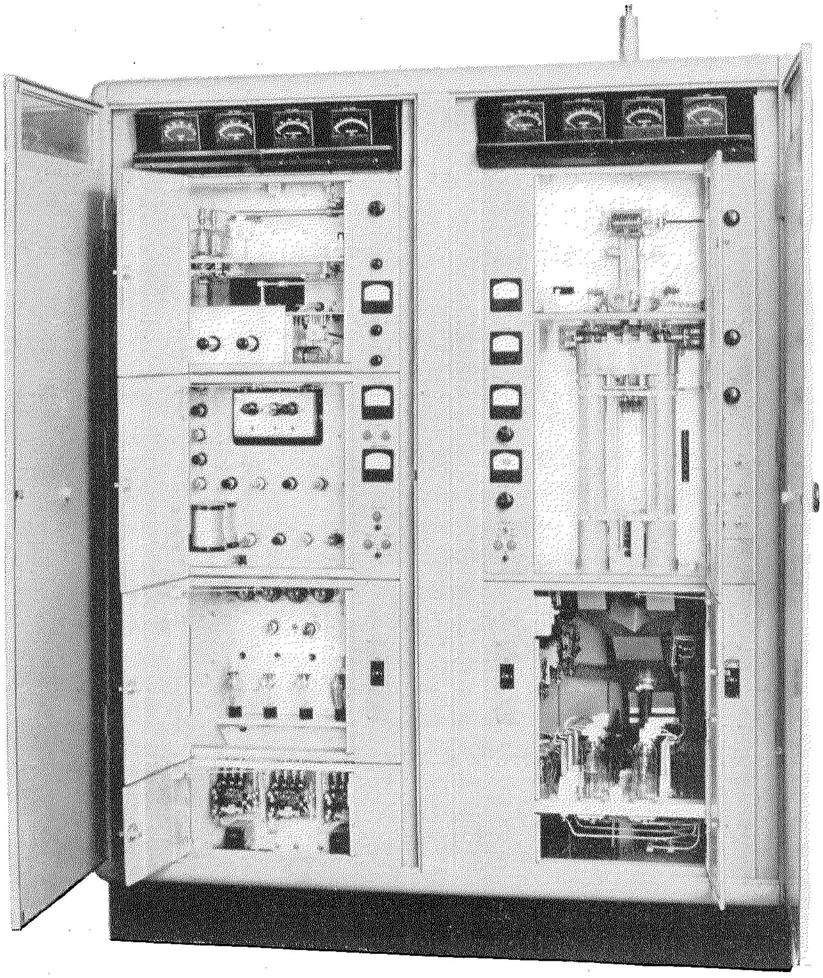


Fig. 4—Response curve of phase detector.

When the transmitter is modulated, instantaneous phase variations are produced in the output of the master oscillator, and these variations are applied through the buffer amplifier and frequency-dividing network to the phase detector. The action of the dividing network is such as to reduce the angle of phase variation by 1/256; consequently, only a relatively small phase change is present at the phase detector. From a mathematical analysis, it can be readily shown that the maximum phase variation at the detector for full modulation is approximately 0.4 radian or 24 degrees.

Under normal operating conditions, the voltages applied to the phase detector have an average phase relationship of approximately 90 degrees, and the detector operates at very near the center of the response curve shown in Fig. 4. The phase variations produced by modulation cause the operating point to be shifted along the response curve, producing corresponding variations in the output voltage of the detector. If these variations were applied to the modulator, they would tend to counteract efforts to modulate the transmitter. In other words, the control system would not only maintain a constant center frequency, but would also try to oppose the desired change in frequency resulting from modulation. To avoid this effect, the output of the phase detector is taken through a 10-cycle low-pass filter, which suppresses all variations except the slow mean changes produced by frequency drift.

In the above description, it was shown that the master oscillator will remain locked in synchronism with the crystal oscillator once it has been brought to its correct frequency. These conditions are fulfilled in normal operation. However, when the transmitter is first turned on, the master oscillator will be out of synchronism with the crystal oscillator. Under these conditions the voltages applied to the phase detector will differ in frequency and have a varying phase relationship. This results in a beat frequency being produced in the output of the phase detector, the beat equaling the difference in frequency of the two applied voltages. This beat-frequency voltage is applied to the modulator producing frequency modulation of the master oscillator, which in turn causes the instantaneous frequency of the master oscillator to be deviated at a rate equal to the beat frequency, and to an extent proportional to the amplitude of the beat. As a result of this action, the frequency of the master oscillator at some instant will pass through its correct frequency. As soon as this occurs, the master oscillator locks into synchronism with the crystal oscillator and the beat frequency disappears. Thereafter, the system operates in the synchronized condition. This lock-in action occurs so rapidly that, when the transmitter is turned on, the master oscillator reaches a synchronized condition



3-kilowatt frequency-modulated transmitter, front view, access doors open.

before the final stages are operative. Thus, the frequency modulation produced by the beat frequency cannot be transmitted.

The frequency of the master oscillator being divided by 256, the beat frequency produced in the phase detector under nonsynchronous conditions, is also effectively reduced by the same ratio. Thus the frequency of the master oscillator can drift as much as 2560 cycles before the beat-note frequency becomes greater than 10 cycles and is suppressed by the 10-cycle low-pass filter. This corresponds to an output-frequency drift greater than 61 kilocycles. Temperature compensation of the tuned circuits in the master oscillator prevents its frequency from drifting outside the region of control.

The number of tuning controls has been kept to a minimum. In the frequency-dividing network, only two of the multivibrator stages are adjustable, the balance of the multivibrator stages requiring no adjustment for operation over the complete range from 88 to 108 megacycles. The number of tuning controls is further reduced by employing a crystal-oscillator circuit of the untuned type. When setting up the modulator for operation, it is merely necessary to insert a crystal of the correct frequency, adjust the two multivibrator stages, and tune three radio-frequency circuits. These circuits are the master-oscillator tank circuit, the modulator plate load circuit, and the buffer-amplifier tank circuit.

A zero-center meter connected to the output of the phase detector indicates the amplitude and polarity of the correcting voltage being applied to the modulator. This meter serves as a carrier-to-crystal phase-comparison meter and tells the operator at a glance wheather or not the master-oscillator tank circuit is correctly tuned.

Standard receiving-type vacuum tubes are employed throughout the modulator and center-frequency-stabilization system. To insure extremely stable and noise-free operation, the filaments of all tubes are operated from direct current, and plate voltages are obtained from a highly regulated power supply.

4. 1- and 3-Kilowatt Transmitters

The 1- and 3-kilowatt transmitters are very similar, the only difference being in their power-supply circuits. The 1-kilowatt transmitter operates from a 230-volt, single-phase supply, while the 3-kilowatt transmitter obtains power from a 230-volt three-phase supply. Also, the high-voltage rectifier employed in the 3-kilowatt transmitter is of larger capacity than that in the 1-kilowatt transmitter and supplies a higher potential to the plate of the final power-amplifier stage. All radio-frequency circuits of the two transmitters are identical, even to the extent of employing the same vacuum tubes.

The accompanying photographs clearly show the unit type of construction employed in all the transmitters. The 1-kilowatt and 3-kilowatt transmitters each consist of two separate units, one being the modulator-exciter unit and the other the power-amplifier unit. The two units are bolted together and, by the addition of side panels, a top, doors, and decorative trim, give the appearance of being housed in a single cabinet.

The operating front of the transmitter proper is enclosed by full-length doors, windows in these doors permitting observation of the principal meters which are located along the top of the transmitter. To provide greatest visibility, the meters have fluorescent scales and pointers and are illuminated with ultraviolet light.

On the left-hand side of the 3-kilowatt unit is the modulator-exciter, and on the right is the power-amplifier unit. The modulator-exciter unit is made up of four vertically mounted chassis. From top to bottom these are: frequency-multiplier and intermediate-amplifier chassis, Frequematic modulator chassis, low-voltage power-supply chassis, and control and contactor chassis. The power-amplifier unit consists of two vertically mounted chassis. The upper is the power-amplifier chassis, and the lower is the high-voltage power-supply chassis.

Access to front-mounted components is obtained through small inner doors on the front of the individual chassis. To protect operating personnel, each door giving access to high voltage is interlocked with the high-voltage power supplies. All vacuum tubes are readily accessible from the front of the transmitter. Doors at the rear of the cabinets provide access to components mounted on the rear of the vertical chassis.

Tuning controls, meters, switches, and supervisory lights are mounted on the panels along each side of the inner doors. This arrangement

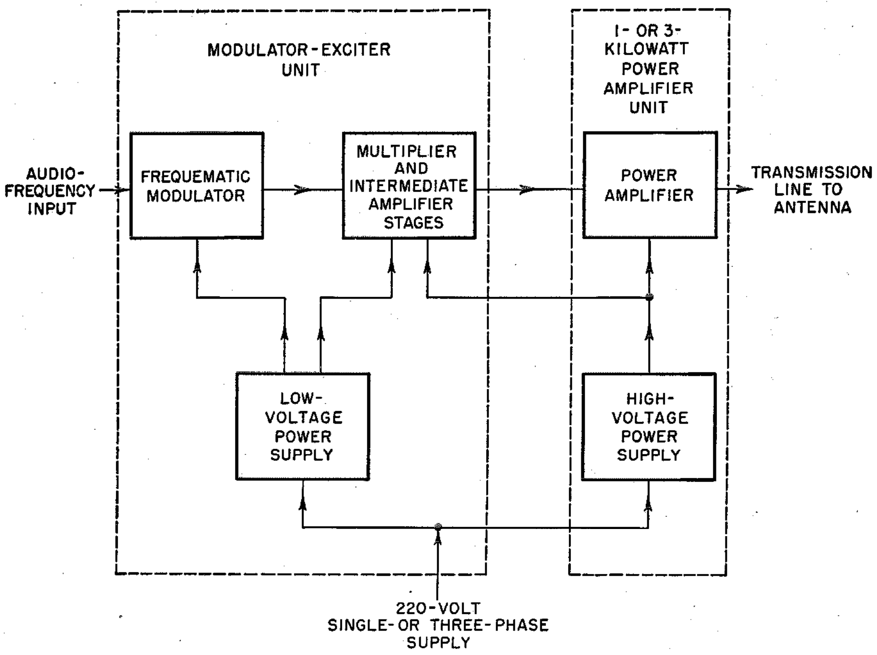

MODULATOR-EXCITER UNIT
1- OR 3-KILOWATT POWER AMPLIFIER UNIT
AUDIO-FREQUENCY INPUT
FREQUEMATIC MODULATOR
MULTIPLIER AND INTERMEDIATE AMPLIFIER STAGES
POWER AMPLIFIER
TRANSMISSION LINE TO ANTENNA
LOW-VOLTAGE POWER SUPPLY
HIGH-VOLTAGE POWER SUPPLY
220-VOLT SINGLE- OR THREE-PHASE SUPPLY

Fig. 5—Block diagram of 1- and 3-kilowatt frequency-modulated transmitters.

permits the transmitter to be completely readjusted from the front. Ventilation and tube-cooling requirements are met by a system of blowers, ductwork, and air filters. In the modulator-exciter unit, the air is directed through a duct that runs up the right-hand side of the unit (as viewed from the rear of the transmitter), openings in one side of the duct providing a horizontal draft across the components. In the 1- and 3-kilowatt power-amplifier units, the air is conveyed through a Y-shaped duct to the transmission-line-type plate inductor and through it to the radiators of each power-amplifier tube. A small amount of air is directed to the base of the rectifier tubes. Air is discharged through louvers at the upper rear and top portion of each assembly. To reduce vibration, the blowers are supported on shock-mounted cradles, and canvas couplings connect the blowers to the air filters and ducts.

The transmitter is normally controlled by a single master "On-Off" switch, sequencing and time-delay cycling being entirely automatic. Individual plate-, bias-, and control-voltage switches are provided, however, to facilitate maintenance and adjustment of the transmitter.

A set of supervisory lights on each unit of the transmitter indicates the condition of the circuits in that unit. These lights are arranged in the order of sequencing and cycling of the transmitter, thus permitting the location of a fault to be determined at a glance. Provision is made for the extension of all control switches and indicating lamps to a remote operating position. Audible warning of plate-current overload and of an excessive temperature condition is given by a buzzer within the equipment.

4.1 MODULATOR

A block diagram of the 1- and 3-kilowatt transmitters is shown in Fig. 5. The audio-frequency input signal is applied directly to the Frequematic modulator. The output of the modulator consists of a frequency-modulated signal having a center frequency between 3.5 and 4.5 megacycles, the exact value depending on the required transmission frequency.

Two plug-in crystals, one operating and one spare, are mounted on the modulator chassis. Each crystal has its own oven and temperature-controlling circuit. The spare crystal may be substituted instantly for the operating one by the flick of a switch.

Filament supply for vacuum tubes in the modulator is obtained from a 12-volt selenium rectifier. Voltage for the plate and screen-grid circuits is provided by the regulated 250-volt output of the low-voltage direct-current power supply.

4.2 FREQUENCY-MULTIPLYING STAGES

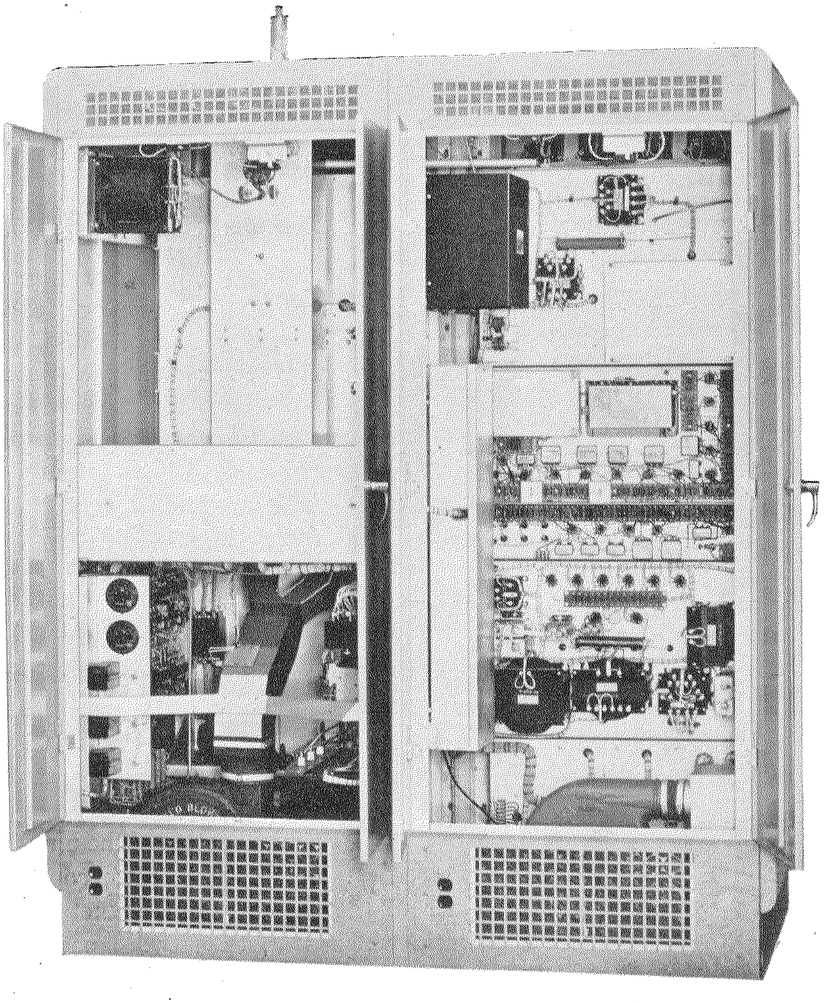
The frequency-multiplying section of the transmitter consists of three doubler stages and one tripler stage. The frequency of the master oscillator is thereby multiplied 24 times to produce the carrier output frequency. The modulating frequency deviation is also multiplied 24 times, thus the maximum frequency swing of the master oscillator, which is about ±3.1 kilocycles is multiplied to ±75 kilocycles, the figure stipulated by the Federal Communications Commission for full modulation.

The three doubler stages employ type 1614 vacuum tubes in essentially standard circuits. The tank circuits of the first and second doublers are loaded by means of resistors to lower the Q and to insure a wide pass-band for the transmission of modulation sidebands. Because of the higher frequency of its plate tank, this type of loading is not required in the third doubler. The output of the third doubler is applied to the grid of a tripler stage which employs a type 815 dual beam power tube with the two sections of the tube used as a push-pull class-C, radio-frequency amplifier. The plate tank circuit is at the output carrier frequency of 88 to 108 megacycles and consists of a variable capacitor and a transmission-line-type inductor, a movable shorting bar providing means of adjusting the effective length of the inductor.

Plate and screen voltage for the multiplying stages are obtained from the low-voltage power supply. The doubler stages have 250 volts on both plate and screen, while the tripler stage has 250 volts on its screen and 400 volts on its plate.

4.3 Buffer Stage

The output of the tripler stage is capacitively coupled to the grids of a buffer stage, which also employs a type 815 dual beam power tube used as a push-pull class-C radio-frequency amplifier. The plate tank circuit is arranged in essentially the same manner as the tripler stage and employs a variable capacitor with a transmission-line-type inductor. Sufficient capacitance for neutralization is obtained by small studs mounted alongside the tube, adjacent to the plate leads, and cross-connected to the opposite grids. Plate and screen voltages are 400 and 250 volts, respectively, being obtained from the low-voltage power supply.



3-kilowatt frequency-modulated transmitter, rear view, doors open.

4.4 The 250-Watt Amplifier Stage

The 250-watt amplifier stage utilizes two Eimac type 4-250A tetrodes as a push-pull class-C radio-frequency amplifier. The plate and grid circuits of this stage are tuned by variable capacitors and transmission-line-type inductors, each inductor being equipped with a sliding shorting-bar for tuning. The grid inductor is inductively coupled to the plate inductor of the buffer amplifier stage. An adjustable loop permits variable inductive coupling to the plate lines for transferring output power to the following stage.

"Neutralization" of this stage is accomplished by a balanced split-stator capacitor having one set of stator plates connected to the screen grid of each of the two tubes and the rotor connected to ground. By adjusting this capacitor, the inductance between the screens and ground is tuned to series resonance, thus minimizing the screen-to-ground impedance and increasing the effectiveness of the screen as a shield between grid and plate.

Plate voltage for this stage is obtained from the high-voltage power supply located in the associated 1- or 3-kilowatt power-amplifier unit; 400 volts for the screen grids is furnished by the low-voltage power supply.

4.5 Low-Voltage Power Supply

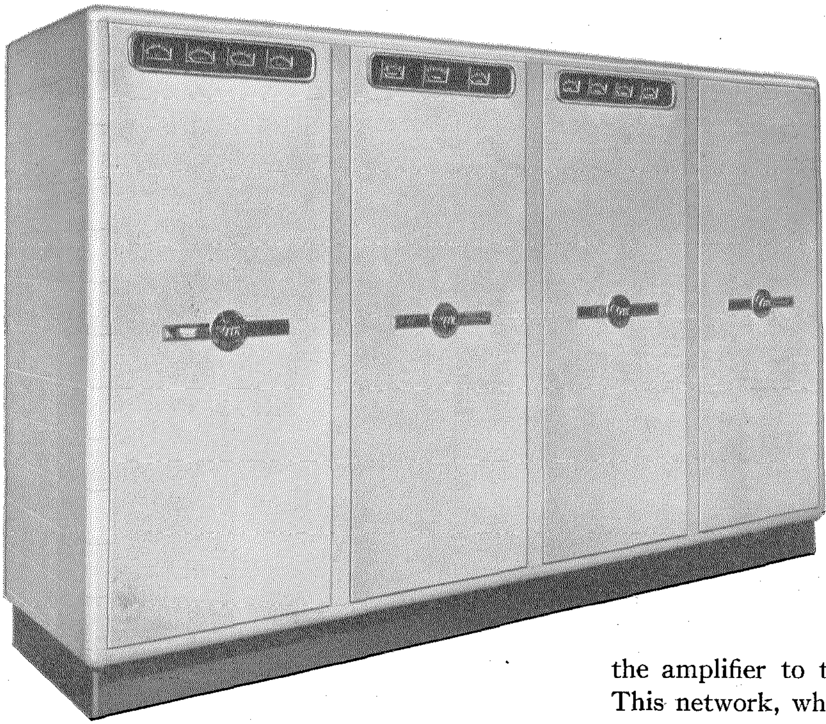
The 250-volt and 400-volt direct-current requirements of the low-power stages are met by a single-phase full-wave rectifier employing four type 866-A mercury-vapor rectifier tubes. The 400-volt supply is obtained from the output of the rectifier through a two-section filter. A 250-volt direct-current potential is obtained by taking the output of the rectifier through an electronic voltage-regulator circuit. This circuit employs six type 6L6G beam power tubes connected in parallel as series regulating tubes. The grids of these tubes are controlled by two stages of direct-current amplification. The output of the voltage-regulator circuit provides an extremely clean direct potential, the variations caused by power-supply ripple and line transients being reduced to as low as 1 to 2 millivolts.

4.6 The 1- and 3-Kilowatt Power Amplifiers

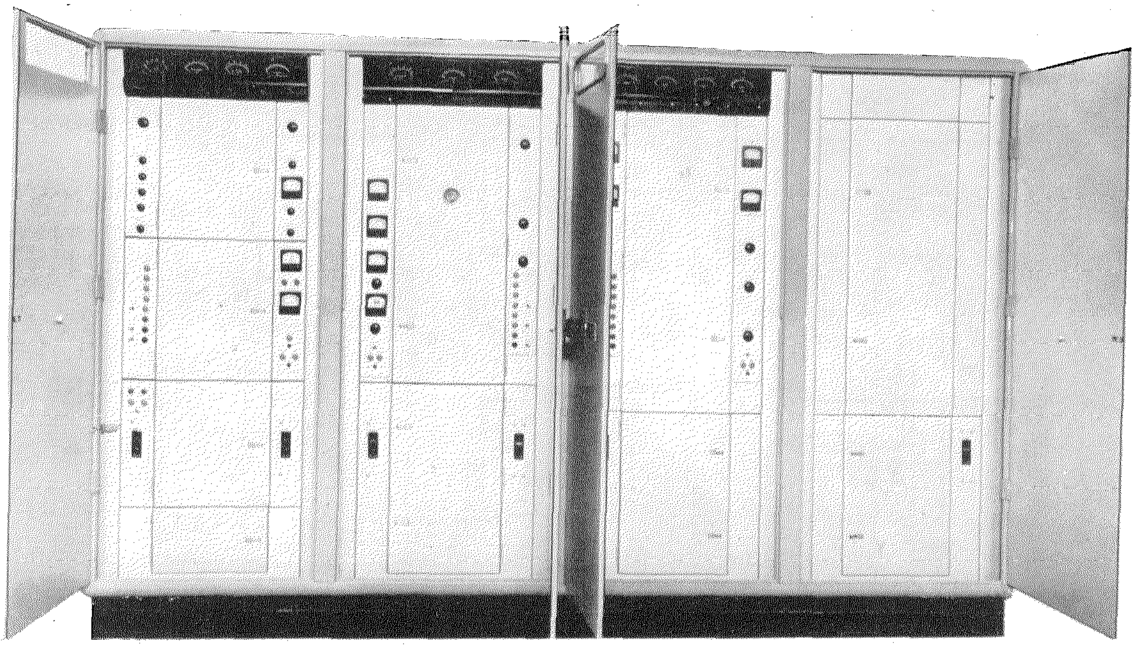
As explained above, the 1- and 3-kilowatt transmitters employ identical power-amplifier stages. These stages employ two Federal type 7C26 vacuum tubes in a push-pull neutralized circuit. The tubes are rated at a maximum plate dissipation of 1 kilowatt each for frequencies up to 150 megacycles.

The output of the 250-watt amplifier is taken through a balanced transmission line to a variable coupling loop located in the power-amplifier stage. This loop transfers power to a transmission-line-type inductor, which, together with a variable capacitor, forms the grid tank circuit of the power amplifier.

The plate tank circuit also consists of a transmission-line-type inductor and a variable capacitor. A dual capacitor, which is part of the plate tank assembly, permits cross-neutralization of the amplifier. A variable loop couples the amplifier to the output matching network. This network, which consists of two capacitors and a transmission-line-type inductor, has two functions. It transforms the single-ended 51-ohm



10-kilowatt frequency-modulated transmitter, front view, doors closed.



10-kilowatt frequency-modulated transmitter, front view, doors open.

impedance of the transmission line, including such variations as may appear with standing-wave ratios as high as 1.75:1, to a balanced impedance into which the coupling loop efficiently transfers the power output; and it adjusts the phase angle of the impedance presented to the transmitter output in such a manner that the plate tuning of the power amplifier is essentially independent of the degree to which the load is coupled to the transmitter. Two coupling loops are provided within the network, one to operate a thermocouple-type radio-frequency ammeter, and the other for a program monitor.

4.7 HIGH-VOLTAGE POWER SUPPLY

The high-voltage power supply for the 1-kilowatt transmitter employs four type 866-A mercury-vapor rectifier tubes operating in two parallel branches as a full-wave rectifier. The primary power for the rectifier is obtained from a 230-volt single-phase source, and the rectifier develops an output of 2000 volts at approximately 1.0 ampere.

In the case of the 3-kilowatt transmitter, the high-voltage power supply employs six type 8008 mercury-vapor tubes operating in a full-wave three-phase circuit. The primary source of power is a 230-volt three-phase supply, and the rectifier delivers 3000 volts at approximately 2 amperes.

A bias supply developing 150 volts at 300 milliamperes is also located on the high-voltage supply chassis. This supply provides fixed grid bias for the final amplifier stage. It employs two type 5R4GY tubes in a full-wave rectifier circuit.

5. 10-Kilowatt Transmitter

The 10-kilowatt transmitter consists of the two units of the 3-kilowatt transmitter to which are added a 10-kilowatt power-amplifier and a 4-kilovolt power-supply. From the accompanying photographs, it can be seen that the general construction of the 10-kilowatt amplifier and its associated power-supply follows the same general lines as that of the 1- and 3-kilowatt transmitters.

Tuning controls, meters, switches, and supervisory lights for the 10-kilowatt amplifier are located on the operating front of the unit. As is the case of the lower-power transmitters, sequencing and cycling of the control circuits are entirely automatic, and operation is controlled by a single master "On-Off" switch. In addition, a special automatic overload reset mechanism is provided for the 4000-volt supply. This mechanism automatically resets the high-voltage-supply overload circuit a fraction of a second after

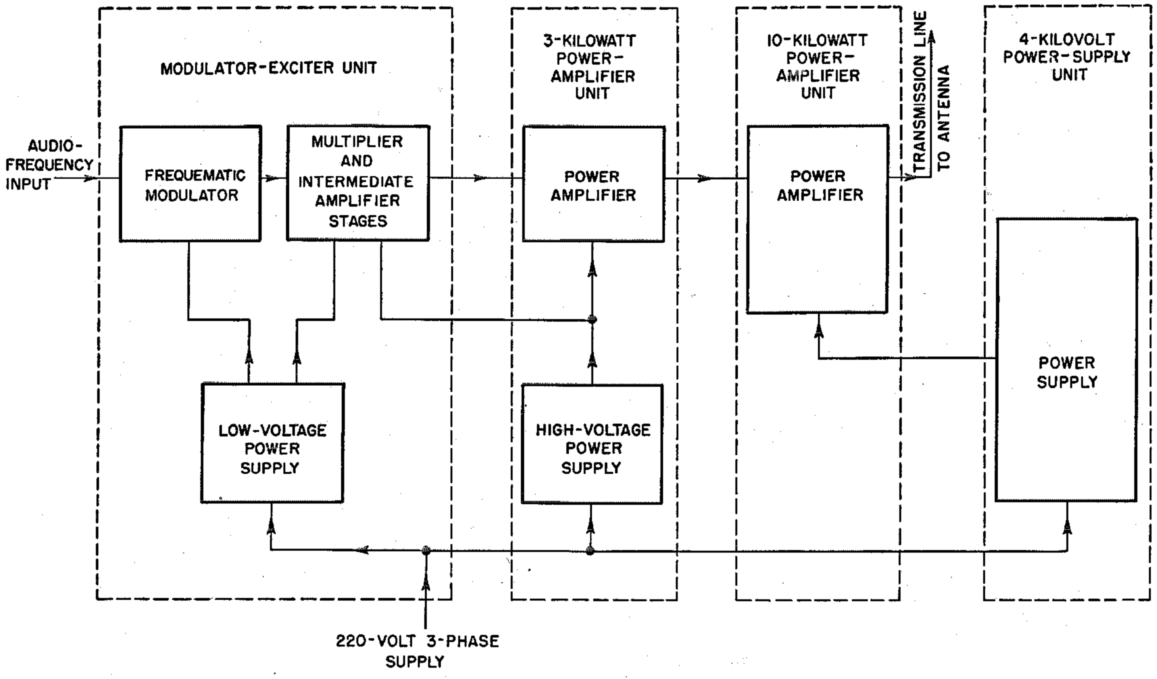

MODULATOR-EXCITER UNIT
3-KILOWATT POWER-AMPLIFIER UNIT
10-KILOWATT POWER-AMPLIFIER UNIT
TRANSMISSION LINE TO ANTENNA
4-KILOVOLT POWER-SUPPLY UNIT
AUDIO-FREQUENCY INPUT
FREQUEMATIC MODULATOR
MULTIPLIER AND INTERMEDIATE AMPLIFIER STAGES
POWER AMPLIFIER
POWER AMPLIFIER
POWER SUPPLY
LOW-VOLTAGE POWER SUPPLY
HIGH-VOLTAGE POWER SUPPLY
220-VOLT 3-PHASE SUPPLY

Fig. 6—Block diagram of 10-kilowatt frequency-modulated transmitter.

it has been tripped. The momentary opening of the high-voltage circuit is usually sufficient to break any temporary flashover, and normal operation can be resumed after a barely noticeable interruption. If the overload is persistent, the overload circuit will again be tripped and reset, this process being repeated three times. After that, if the overload is still present, the power-supply circuit remains open and an alarm buzzer is sounded.

A block diagram of this transmitter is shown in Fig. 6. The 3-kilowatt transmitter drives the 10-kilowatt amplifier. Under these conditions, the output matching network in the 3-kilowatt amplifier is not required, and the output coupling loop of the 3-kilowatt stage is connected through a balanced transmission line to the 10-kilowatt amplifier.

The 10-kilowatt power amplifier employs two Federal type 7C27 tubes in a push-pull grounded-grid circuit. Considerably greater driving power is required with the grounded-grid circuit than with a conventional grounded-filament circuit. However, the major portion of this power is added to that in the output circuit, thus increasing the power output of the amplifier.

The output of the 3-kilowatt amplifier is fed through a coupling loop to the tuned filament circuit of the 10-kilowatt amplifier, which consists of a transmission-line-type inductor and a variable capacitor. The inductor is constructed of hollow tubing, and the filament leads pass through the center of this inductor.

The plate tank circuit is essentially the same as for the 3-kilowatt amplifier. However, the components are increased slightly in size to accommodate the larger tubes and the higher power dissipation.

The output of the amplifier is taken from a coupling loop and fed through a matching network and coaxial line to the antenna. The matching network is identical to that employed in the 3-kilowatt transmitter.

A low-voltage bias supply is located below the amplifier chassis. It employs two type 5R4GY tubes and supplies 200 volts of negative bias to the grids of the amplifier tubes.

Plate voltage for the amplifier is obtained from the 4-kilovolt rectifier located in the high-voltage power-supply unit. This rectifier employs six type 8008 mercury-vapor tubes in a three-phase full-wave circuit. The output current capabilities of this rectifier have been materially increased by employing a circuit in which the plate voltage is applied in quadrature to the rectifier filament voltage. Use of this circuit has made it possible to employ relatively low-cost rectifier tubes resulting in an appreciable saving. The power-supply unit also houses the relays and contactors which form a part of the control and protective circuits.

A blower located on the floor of the amplifier unit supplies air for cooling components in both the amplifier and the power-supply units. The air is drawn in through a filter located at the rear of the unit and flows through a Y-shaped duct to the plate tank of the amplifier. The air then flows through the tank inductor assembly to the plate cooling radiators of the amplifier tubes. Two smaller ducts branch off from the base of the Y to direct air into the power-supply unit. One of these ducts supplies air to the lower part of the unit for cooling the transformers, while the other carries air to the bases of the type 8008 rectifier tubes.

Theory and Design of the Reflectometer

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REFLECTION coefficient and standing-wave ratio are related quantities, each being a measure of impedance mismatch in a transmission system. A high reflection coefficient, caused either by mismatches or discontinuities in the line, will result in low power-transmission efficiency; an amount of power proportional to the degree of mismatch will be reflected back toward the generator and thus produce standing waves.

Simple, accurate devices are available for testing open-wire lines operating at low frequencies. The "reflectometer" has been designed for measurement of standing waves, reflection coefficient, and power transfer on coaxial lines operated at 1000 megacycles per second and below.

* * *

The reflectometer is inserted into a coaxial line at a desired location for measurement of reflection coefficient, standing-wave ratio, and power transfer in the line. The instrument consists of a short section of coaxial line containing a small loop coplanar with the inner conductor. The loop is connected through a resistor to the outer conductor, and this resistor is capacitively coupled to the inner conductor. The voltage appearing across the series arrangement of loop and resistor is measured when the voltage across the resistor and the voltage induced in the loop are aiding, and again when they are in opposition to each other. These two readings are obtained by rotating the loop through 180 degrees. As will be shown below, the readings may be used to determine the amount of mismatch and the power carried by the line. Operation is substantially independent of load impedance and meter impedance at any frequency within the useful range of the instrument. In the present models, voltage is measured by a diode voltmeter. A reflectometer with a fixed loop assembly is useful for continuous monitoring of line power. It can also be used to measure standing-wave ratio by reversing the line connections, which is equivalent to reversing the loop. The design shown operates satisfactorily in the region of 1000 megacycles. A reflectometer unit with a rotating loop assembly has been designed for use in the 600-megacycle region.

1. Basic Theory

If a resistance and capacitance connected in series are placed between the inner and outer conductors of a coaxial line, as drawn in Fig. 1, the voltage e_1 across R will be proportional to E the voltage between the conductors.

If a loop is placed between the inner and outer conductors, as in Fig. 2, the voltage e_2 induced in the loop is proportional to the radio-frequency current I in the inner conductor.

In practice, e_1 and e_2 are simultaneously obtained by using the arrangement in Fig. 3. The necessary capacitive coupling between the resistor and inner conductor is provided by the proximity of the bottom of the resistor (and the connected loop wire) to the conductor.

Since all dimensions are chosen to be much smaller than λ, the distributed parameters can be approximately replaced by lumped impedances as shown in Fig. 4, where

C = coupling capacitance,
M = mutual inductance between the loop and the reflectometer line,
E = voltage between conductors, and
I = current flowing in the line.

The voltage across R will then be given by

$$\frac{RE}{R-jX_c}=\frac{RE(R+jX_c)}{R^2+X_c^2},$$

where $X_c=1/\omega C$, and $\omega=2\pi f$. Let $R\ll X_c$. The voltage across R will be given by

$$e_1\simeq\frac{RE}{X_c}=j\omega CRE.$$

The voltage induced in the mutual inductance M is

$$e_2 = j\omega MI.$$

The total electromotive force induced is therefore

$$e = j\omega(CRE + MI). \qquad (1)$$

In this equation, M may have positive or negative values.

The current and the voltage in the line may each be considered to consist of a wave traveling forward, and a reflected wave:

$$\left.\begin{aligned} E &= E_f + E_r, \\ I &= I_f + I_r, \end{aligned}\right\} \qquad (2)$$

where the subscript f denotes the forward wave, and r denotes the reflected wave. It can be shown that

$$\left.\begin{aligned} E_f &= I_f Z_0, \\ -E_r &= I_r Z_0, \end{aligned}\right\} \qquad (3)$$

where Z_0 is the characteristic impedance of the reflectometer line. It is further known that the reflection coefficient Δ is given by

$$\Delta = E_r/E_f = -I_r/I_f. \qquad (4)$$

Substituting (2), (4), and (3) into (1), we have

$$\begin{aligned} e &= j\omega[RC(E_f + E_r) + M(I_f + I_r)], \\ &= j\omega[RC(E_f + E_f\Delta) + M(I_f - I_f\Delta)], \\ &= j\omega\left[RC(E_f + E_f\Delta) + M\left(\frac{E_f}{Z_0} - \frac{E_f\Delta}{Z_0}\right)\right], \\ &= j\omega E_f\left[RC(1+\Delta) + \frac{M}{Z_0}(1-\Delta)\right]. \end{aligned}$$

Let R, C, M and Z_0 be chosen so that

$$RC = M/Z_0 = A, \qquad (5)$$

where A is a given constant. Let e_p be the electromotive force when M is positive, and e_n the electromotive force when M is negative. The reversal of the mutual inductance is accomplished by turning the loop through 180 degrees, so as to reverse the direction of flux linkage in it. Then,

$$\left.\begin{aligned} e_p &= j\omega E_f[A(1+\Delta) + A(1-\Delta)] \\ &= 2j\omega E_f A, \end{aligned}\right\} \qquad (6)$$

and

$$\left.\begin{aligned} e_n &= j\omega E_f A(1 + \Delta - 1 + \Delta) \\ &= 2j\omega E_f A\Delta. \end{aligned}\right\} \qquad (7)$$

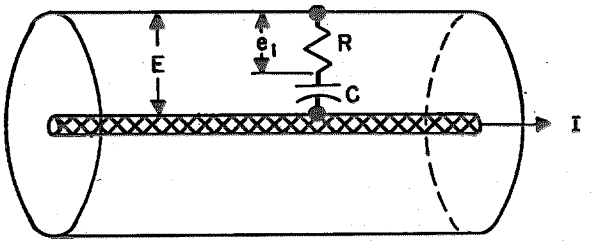


Fig. 1—Voltage relationships for a series resistance-capacitance combination placed between the conductors of a coaxial line; e_1 is proportional to E.

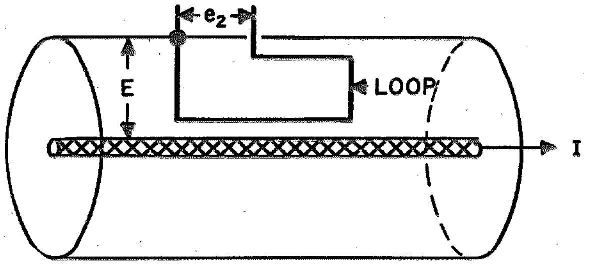


Fig. 2—A loop coupled to the inner conductor of a coaxial line will give a voltage output proportional to the current flowing in the line.

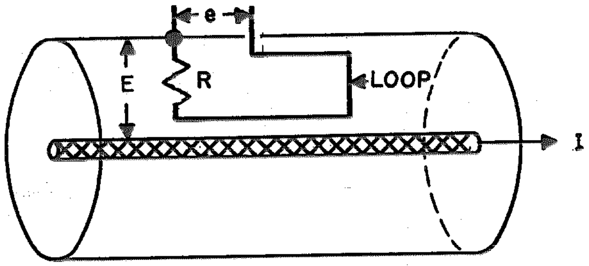


Fig. 3—Schematic representation of the reflectometer, showing loop and resistance. Capacitance is provided by the proximity of the arrangement to the inner conductor.

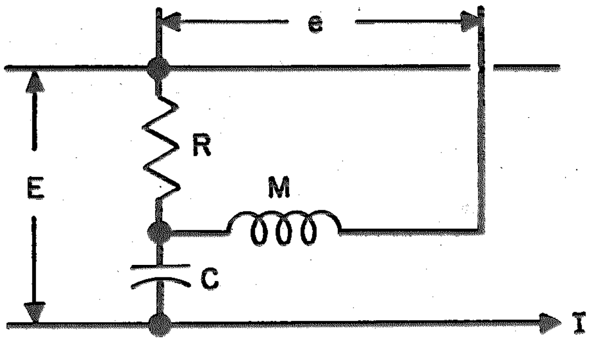


Fig. 4—Equivalent circuit of reflectometer.

Dividing (7) by (6), we have

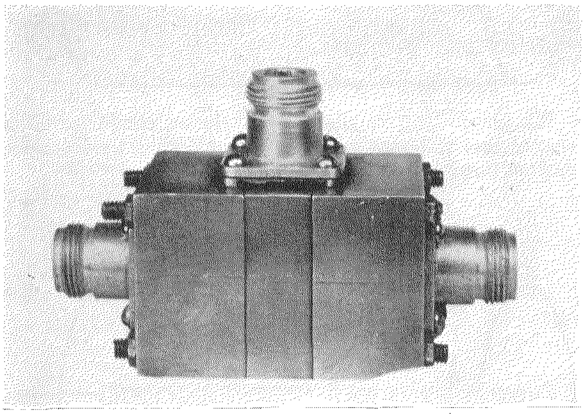
$$\Delta = e_n/e_p$$

$$|\Delta| = \left|\frac{e_n}{e_p}\right| = \frac{|e_n|}{|e_p|}. \tag{8}$$

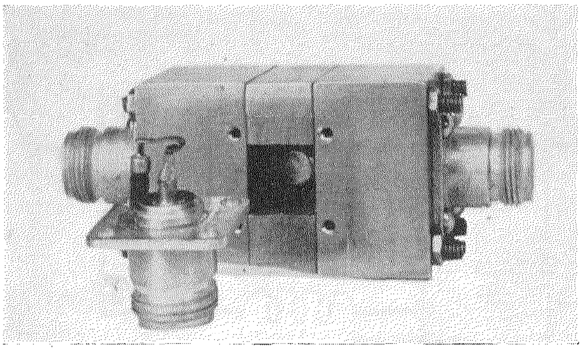
If the line has no attenuation and characteristic impedance Z_0 then the standing-wave ratio q is given by

$$q = \frac{1+|\Delta|}{1-|\Delta|} = \frac{1+|e_n/e_p|}{1-|e_n/e_p|},$$

$$q = \frac{|e_p|+|e_n|}{|e_p|-|e_n|}. \tag{9}$$



Reflectometer with fixed loop assembly for 1000-megacycle range. Power flow is directly through the pickup, with probe output connector at the top. Length of block is 2¼ inches.



Disassembled view of fixed-loop reflectometer. Note capacitance disk attached to inner conductor of line section, and loop and resistor attached to the dismounted output probe connector.

The reflectometer may be used to measure power P as may be demonstrated in the following manner:

$$P = E \cdot I, \text{ where } E \cdot I = EI \cos\theta.$$

From (2),

$$P = E \cdot I = (E_f + E_r) \cdot (I_f + I_r),$$
$$= E_f \cdot I_f + E_r \cdot I_f + E_r \cdot I_r + E_f \cdot I_r.$$

But from (4), $E_r \cdot I_f = -E_f \cdot I_r$, so that

$$P = E_f \cdot I_f + E_r \cdot I_r. \tag{10}$$

Substituting from (3) and (4),

$$P = \frac{E_f \cdot E_f}{Z_0} - \frac{E_r \cdot E_r}{Z_0} = \frac{|E_f|^2}{Z_0} - \frac{|E_r|^2}{Z_0}$$

$$= \frac{|E_f|^2}{Z_0}(1-|\Delta|^2).$$

By (8), $|\Delta| = |e_n/e_p|$. Then,

$$P = \frac{|E_f|^2}{Z_0}\left(1 - \left|\frac{e_n}{e_p}\right|^2\right), \tag{11}$$

$$= \frac{|E_f|^2}{Z_0|e_p|^2}(|e_p|^2 - |e_n|^2).$$

From (6), $\left|\frac{E_f}{e_p}\right| = \frac{1}{2\omega A}$, therefore

$$P = \frac{|e_p|^2 - |e_n|^2}{4\omega^2 A^2 Z_0} = \frac{c(|e_p|^2 - |e_n|^2)}{\omega^2}, \tag{12}$$

where $c = 4A^2 Z_0 = 4M^2/Z_0$, and depends only on the circuit constants of the reflectometer.

The value of c may be obtained by calculation when the circuit constants are known. R and Z_0 can be accurately determined, and M is rather simply calculable from the geometry of the reflectometer. The value of the capacitance may be adjusted to agree with the design condition $C = A/R = M/Z_0R$, by using the reflectometer to measure the standing-wave ratio in a line, and matching the observed value to that given by slotted-line measurements. Having determined the appropriate circuit constants, (12) shows that at a known frequency, and for any load impedance, one may use the reflectometer to give an absolute value of power without the necessity of calibrating against a known power, which is usually difficult to perform.

2. *Effect of Meter Impedance*

The foregoing discussion has implied the use of a meter having infinite impedance for measurement of the induced voltages. Since this will obviously not be true in practice, the effect of the meter impedance Z_m must be investigated. It is desired that the form of the expressions for reflection coefficient and power do not change when the measured voltages are substituted for the induced electromotive forces.

To determine the effect of the meter resistance, Thevenin's theorem may be used. According to this theorem, the pickup will act as a generator producing the equivalent of the open-circuited electromotive force as determined above, and having an internal impedance equal to the impedance observed when looking back from the load (meter circuit) to the pickup.

With reference to Fig. 5,

X_L = inductive reactance of the loop,
X_c = capacitative reactance between the loop and the inner conductor,
Z_t = total impedance between the inner and the outer conductor of the line, and
Z_r = reflected impedance in the loop due to mutual induction between the loop and the line.

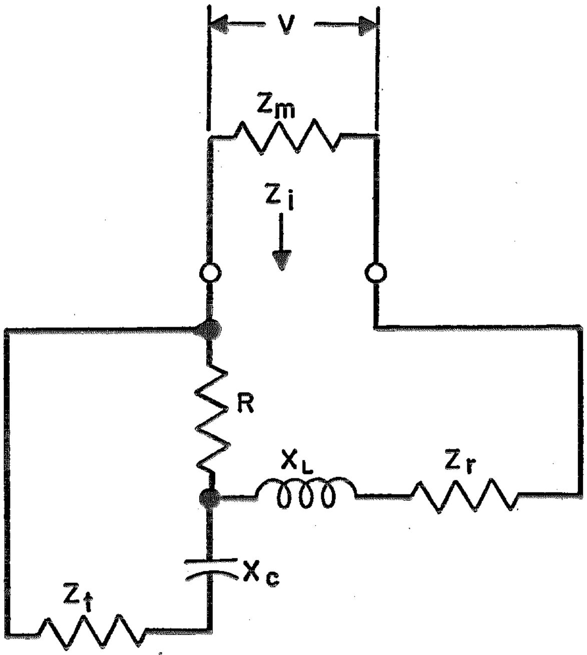
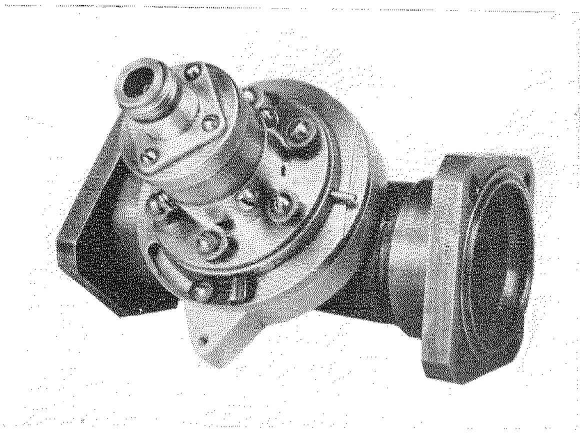
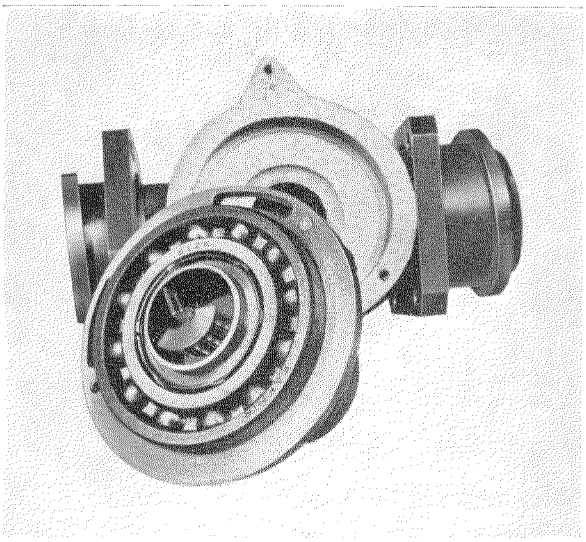


Fig. 5—Reflectometer circuit indicating effect of meter impedance.



Rotating-loop reflectometer for the 600-megacycle range. The probe output connector and rotating mount are attached to the side of a section of standard 1⅝-inch coaxial line.



Disassembled view of rotating-loop reflectometer. The probe mount rotates on ball bearings. The loop is visible in the center of the mount. The resistors are partly hidden from view by the capacitance plate attached to the loop.

The reflected impedance Z_r is given by

$$Z_r = \omega^2 M^2 / Z_t\,, \tag{13}$$

and the internal impedance of the pickup Z_i as viewed from the meter, is therefore given by

$$Z_i = \|R,\, Z_t - jX_c\| + j\omega L + \frac{\omega^2 M^2}{Z_t}\,. \tag{14}$$

The notation $\|Z_1, Z_2, \ldots . Z_n\|$ will be used to designate the impedance of $Z_1, Z_2, \ldots . Z_n$ in parallel. Denote by v_p the voltage observed when M is positive and by v_n the observed voltage when M is negative. Then

$$v_p = e_p\, Z_m/(Z_m+Z_i), \tag{15}$$

and

$$v_n = e_n\, Z_m/(Z_m+Z_i). \tag{16}$$

Equation (14) shows that Z_i is independent of the sign of M, and hence is the same for each position of the loop. Dividing (16) by (15) and substituting in (8), we obtain

$$\Delta = \frac{e_n}{e_p} = \frac{v_n}{v_p}, \quad \text{and} \quad |\Delta| = \left|\frac{v_n}{v_p}\right| = \frac{|v_n|}{|v_p|}, \tag{17}$$

which corresponds to (8). This result is independent of frequency, of the meter impedance, and of its response-frequency characteristics.

The power traversing the line has been shown to be

$$P = \frac{|E_f|^2}{Z_0}\left(1 - \left|\frac{e_n}{e_p}\right|^2\right).$$

From (17),

$$P = \frac{|E_f|^2}{Z_0}\left(1 - \left|\frac{v_n}{v_p}\right|^2\right). \tag{18}$$

Using (6) and (5) it is seen that

$$\begin{aligned} \frac{|E_f|^2}{Z_0} &= \frac{1}{Z_0}\left|\frac{e_p^2}{2j\omega A}\right| = \frac{1}{Z_0}\left|\frac{e_p Z_0}{2j\omega M}\right|^2 \\ &= Z_0 |e_p|^2/4\omega^2 M^2. \end{aligned} \tag{19}$$

Also from (14) and (15),

$$|e_p|^2 = |v_p|^2 \left|\frac{Z_m + \|R, Z_t - jX_c\| + j\omega L + \omega^2 M^2/Z_t}{Z_m}\right|^2.$$

On substituting into (18), it is seen that the power will be proportional to the product of $|v_p|^2$ and constants which are independent of the sign of M.

It is necessary to examine the dependence of the reading on the load impedance, because it affects Z_t, and hence v. From our previous assumption that $X_c \gg R$, and a further condition that R will be of the same order of magnitude as Z_t, $\|R, Z_t - jX_c\| \simeq R$. Furthermore, to avoid large mismatching, Z_t will be of the same order of magnitude as Z_0. It will be shown below that $\omega^2 M^2/Z_0$ is negligible in comparison with R, and that $\omega L \ll R$. With these approximations, (14) reduces to

$$Z_i \simeq R,$$

and

$$\frac{e_p}{v_p} = \frac{Z_m + R}{Z_m} = k,$$

where k is a constant for a given pickup and meter.

From (18) and (19), we then have

$$\begin{aligned} P &= \frac{Z_0 k^2 |v_p|^2}{4\omega^2 M^2}\left(1 - \left|\frac{v_n}{v_p}\right|^2\right) \\ &= c' \, \frac{|v_p|^2 - |v_n|^2}{\omega^2}, \end{aligned} \tag{21}$$

where $c' = \dfrac{Z_0(Z_m+R)^2}{4M^2 Z_m^2}$, and is also a constant.

Equation (21) is in the same form as (12), which holds for the case of infinite meter impedance. For a finite meter impedance, therefore, an absolute determination of the power can also be made from a knowledge of the circuit constants in the reflectometer.

3. *Frequency Limits of Reflectometer*

The most stringent frequency limitation on the operation of the reflectometer is that the wavelength be very much larger than the dimensions of the pickup. This corresponds roughly to a frequency limit of 1000 megacycles. Other factors which affect the operation of the reflectometer as a wattmeter are shown below to be relatively small, and to be independent of frequency when suitably designed.

For the meter to measure power independently of the load impedance, it is necessary that

$$\left|\frac{\omega^2 M^2}{Z_t}\right| \simeq \frac{\omega^2 M^2}{Z_0} \ll |Z_m + R + j\omega L|.$$

To make this true it is sufficient that $|\omega^2 M^2/Z_0| \ll |R|$. By choosing typical values for the circuit constants, the validity of this assumption can be tested: Let

$Z_0 = 50$ ohms,
$E = 100$ volts $= E_f$,
$e_p = 10$ volts, and
$R = 50$ ohms.

A constant value has been chosen for e_p, since in practice the circuit constants of reflectometers operating in widely differing ranges will be adjusted to provide optimum accuracy and the same input voltage for a full-scale meter reading in each range.

From (5) and (6),

$$A = \frac{M}{Z_0} = RC = \frac{e}{2E_f\omega} = \frac{10 \text{ volts}}{2(100 \text{ volts})\omega}$$

$$= \frac{0.05}{\omega} \text{ seconds.}$$

$$M = AZ = 2.5/\omega \text{ henries.}$$

$$\frac{\omega^2 M^2}{Z_0} = \frac{\omega^2 (2.5/\omega)^2}{50} = 0.125 \text{ ohm.}$$

This is negligible compared to R (=50 ohms).

The value of the $j\omega L$ term in (20) may also be calculated

$$j\omega L = 2j\omega M = 2j\omega \frac{(2.5)}{\omega} = 5.0j \text{ ohms.}$$

These results confirm that at all frequencies $|Z_i| \approx R$ to a high degree of approximation (about 1 percent error for the assumption made).

A second factor which must be considered is the validity of the assumption that

$$|R| \ll |jX_c|,$$

or

$$R \ll 1/\omega C.$$

From the previous calculations, $C = A/R = \frac{0.05/\omega}{50} = 0.001/\omega$. Therefore, $X_c = 1/\omega C = 1000$ ohms, which is much larger than 50 ohms: The validity of the assumption is thus established.

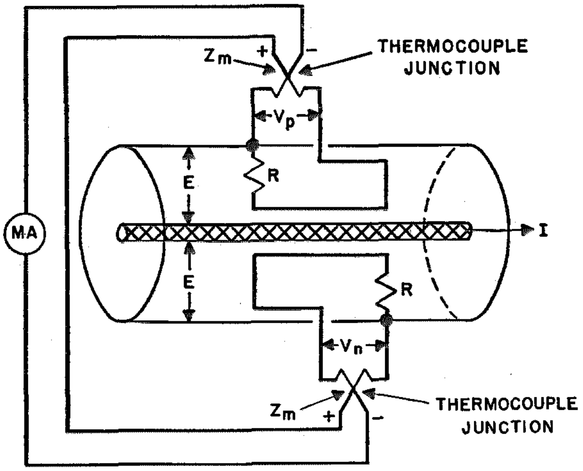


Fig. 6—Directional-pickup principle used in conjunction with thermocouples for measurement of radio-frequency power.

It is possible, therefore, to operate the reflectometer at all frequencies for which the dimensions of the reflectometer loop are not an appreciable fraction of a wavelength. Negligible error will be introduced by other frequency effects.

4. *Applications of Reflectometer*

When the reflectometer is used to measure the reflection coefficient and the standing-wave ratio, the value obtained for these quantities depends only on the ratio of two measured voltages. As was shown above, the results are therefore valid irrespective of variations in load impedance, meter impedance, response, and frequency, provided that the wavelength is greater than about 50 times the effective length of the pickup, i.e., for frequencies less than about 1000 megacycles with the pickups now used. A simple method of measuring Δ would be to use a meter with variable sensitivity, and adjust it so that v_p gives a full-scale deflection. The fraction of full-scale deflection observed when the pickup is rotated through 180 degrees will equal the reflection coefficient. If two identical pickups were used, placed so that M has opposite signs, a ratiometer reading would give Δ directly.

The requirements for use of the reflectometer as a radio-frequency wattmeter are more stringent, since the absolute value of E must be determined. A limit will again be imposed by the size of the pickup. Another requirement for a wide-band wattmeter is the necessity that the meter have a flat frequency response. The reflectometer may be designed so that it operates as a true wattmeter over any desired frequency range, but operation of a given reflectometer at a frequency very much higher than the one for which it was designed will lead to some error.

A further modification would be the use of thermocouples, which have a response proportional to v^2. If two thermocouples in series opposition are used, each connected to one of a pair of pickups which differ only in the sign of M and which are placed in the same section of one line, the resultant reading at a fixed frequency will be directly proportional to the radio-frequency power. Fig. 6 shows such a design. Since very-high-frequency thermocouples are not available, the maximum usable frequency would be reduced to a few hundred megacycles. The maximum frequency is greatly increased by replacing the thermocouple with crystals operating in the square-law region.

Some simplification of design is achieved if a common resistor R is used in the above case, as shown in Fig. 7. The equivalent circuit is given in Fig. 8, where it is assumed that $R \ll X_c$, and $\|R, Z-jX_c\| = R$. Here e_1 and e_2 are the electromotive forces induced, and Z_1 and Z_2 are the meter-circuit resistances for each of the two loops, respectively.

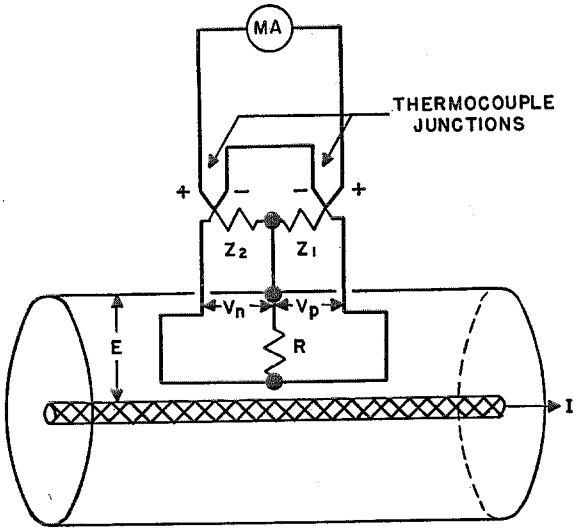


Fig. 7—Simplification of circuit shown in Fig. 6 by use of a common resistor.

From the principle of superposition,

$$v_p = \frac{e_1 Z_1}{Z_1 + \|R, Z_2, X_c\|} + \frac{E\|R, Z_1, Z_2\|}{X_c} + \frac{e_2\|R, Z_1\|}{Z_2 + \|R, Z_1, X_c\|}.$$

Since

$$R \ll X_c, \quad \|R, Z_2, X_c\| \simeq \|R, Z_2\|,$$

and

$$\|R, Z_1, X_c\| \simeq \|R, Z_1\|,$$

then

$$v_p = \frac{e_1 Z_1}{Z_1 + \|R, Z_2\|} + \frac{E\|R, Z_1, Z_2\|}{X_c} - \frac{e_2\|R, Z_1\|}{Z_2 + \|R, Z_1\|}.$$

Similarly,

$$v_n = \frac{-e_1\|R, Z_2\|}{Z_1 + \|R, Z_2\|} + \frac{E\|R, Z_1, Z_2\|}{X_c} - \frac{e_2 Z_2}{Z_2 + \|R, Z_1\|}.$$

If $Z_1 = Z_2 = Z$, and $M_1 = M_2 = M$, so that $e_1 = e_2 = e$, then

$$v_p = \frac{eZ}{Z + \|R, Z\|} + \frac{E\|R, Z/2\|}{X_c} + \frac{e\|R, Z\|}{Z + \|R, Z\|}$$
$$= \frac{E\|R, Z/2\|}{X_c} + e$$
$$= j\omega(\|R, Z/2\|CE + MI),$$

and

$$v_n = j\omega(\|R, Z/2\|CE - MI).$$

By comparison of these basic equations with those of the simple case (1), it can be seen that a similar set of equations for Δ and P will be obtained if the following design condition is used:

$$A = \frac{M}{Z_0} = C\|R, Z/2\| = \frac{RZC}{Z + 2R}.$$

In that case

$$|\Delta| = \left|\frac{v_n}{v_p}\right|,$$

and

$$P = \frac{Z_0}{4M^2\omega^2}(|v_p|^2 - |v_n|^2).$$

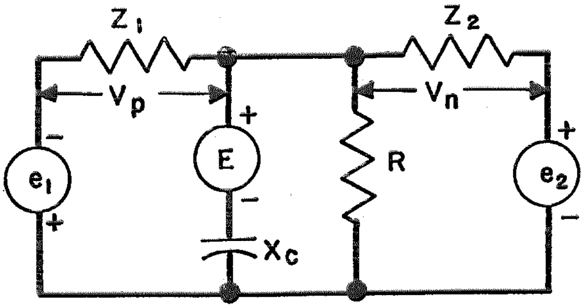


Fig. 8—Equivalent circuit of arrangement shown in Fig. 7.

5. *Conclusions*

The reflectometer is more compact than the present design of instruments which it replaces. It provides a measure of Δ and standing-wave ratio using a measuring line which extends only over a small fraction of a wavelength, in contrast to slotted-line measurements which require at least quarter-wavelength lines. Furthermore, in contrast to present forms of radio-frequency wattmeters employed at high radio frequencies, it does not depend on resonant circuits, and hence is much less sensitive to frequency variation.

Cathode-Design Procedure for Electron-Beam Tubes

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Editor's Note: The results of a research project carried on at Stanford University during 1941 and 1942 under the sponsorship of the International Telephone and Telegraph Corporation were published in part in a book entitled The Production and Control of Electron Beams *by Karl Spangenberg, Lester M. Field, and Robert Helm. This limited edition was designated as "Secret" during the war, but is now declassified. This paper is based on parts of the book, which will not be reprinted in its entirety.*

DESIGN problems of single-potential cathode structures to produce convergent beams of specified size and current and of uniform current density have been investigated, and design procedures developed. All such structures consist of an electrode at cathode potential surrounding the cathode, and an accelerating electrode with an aperture through which the major portion of the beam passes. A typical structure is shown in Fig. 1. It is, of course, desirable to pass as high a percentage as possible of beam current through the aperture.

In general, the problem of calculating electron paths under conditions of space-charge-limited current flow between electrodes of arbitrary shape is difficult, if not impossible. There are well-known laws, however, governing space-charge-limited current flow for certain specific cases, such as current flow between parallel planes, concentric cylinders, and concentric spheres.[1,2,3] For all of these cases, the position of any electron can be stated in terms of only one coordinate.

J. R. Pierce proposed that to produce a uniformly convergent beam it was merely necessary to consider the beam to be a cone cut from inward radial current flow between concentric spheres, and then to shape the cathode electrode and the accelerating electrode to give the same electric field at the edge of the beam as did the rest of the spherical flow that had been cut away.[3] Since the beam is a cone cut from space-charge-limited current flow between concentric spheres, both the current-voltage relations and the

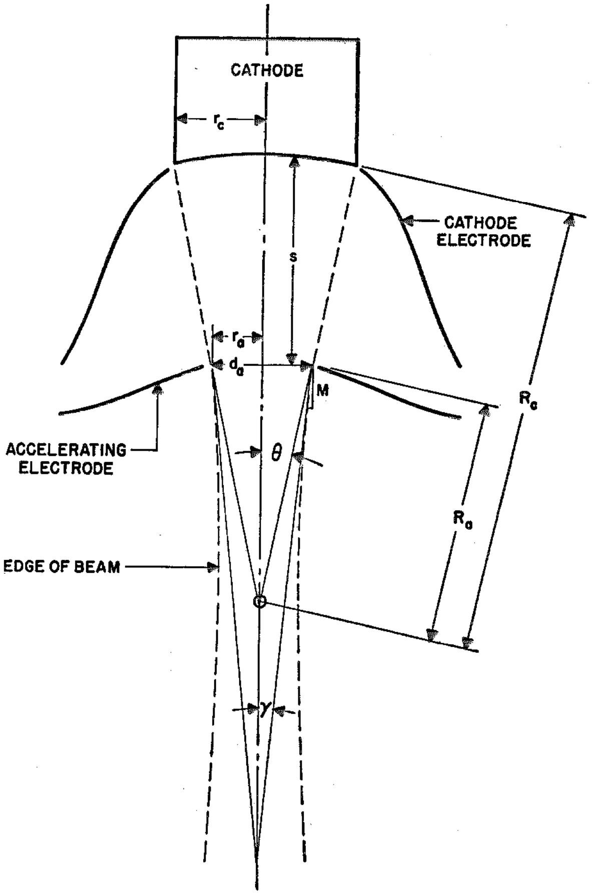

Fig. 1—Cross-section of typical cathode structure, showing dimensions involved in determination of beam shape. The electrode shapes are not functions of absolute size, but rather of ratios between dimensions.

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[1] C. D. Child, "Discharge from Hot CaO," *Physical Review*, Series 2, v. 32, pp. 492–551; 1911.

[2] I. Langmuir and K. B. Blodgett, "Currents Limited by Space Charge Between Concentric Spheres," *Physical Review*, Series 2, v. 24, pp. 49–59; 1924.

[3] J. R. Pierce, "Rectilinear Electron Flow in Beams," *Journal of Applied Physics*, v. II, pp. 548–554; August, 1940.

variation of potential with radius, which determines the electrode shapes, are known.

This design procedure allows us to calculate all the operating characteristics of the cathode structure. The current-voltage relationship, the angle of convergence of the beam (both in the cathode-accelerating electrode space and also in the space behind the accelerating electrode), the cathode-current density, and the size and current density of the beam as it leaves the accelerating electrode are all known quantities.

The relations between current, voltage, and the geometry of the structure have been reduced to the chart of Fig. 2. The dimensional quantities involved are shown in Fig. 1, which is a cross section of a typical cathode structure. The characteristics of these structures are, as would be expected, functions only of ratios of dimensions, and are independent of the actual size of the structure. For this reason any convenient system of dimensions may be used, and the absolute size of the structure may be made such that the full emission density will be drawn from the type of emitting surface selected. For the chart of

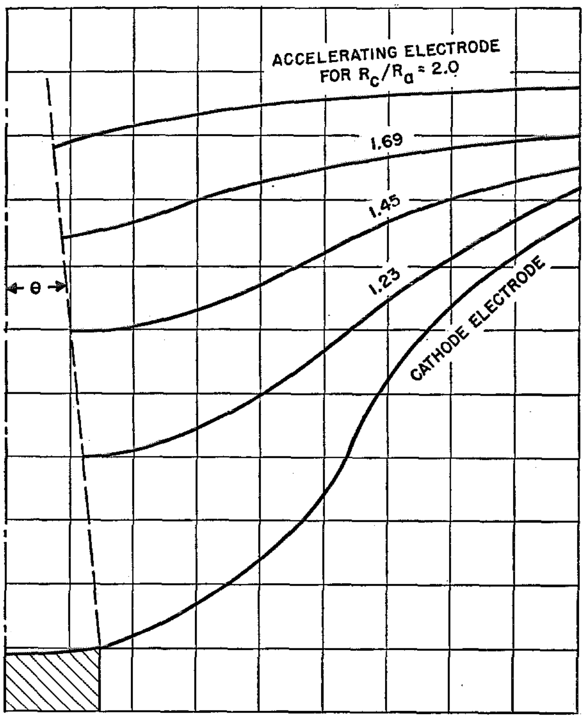

Fig. 3—Shape and placement of accelerating electrode for indicated values of R_c/R_a, where $\theta = 5$ degrees.

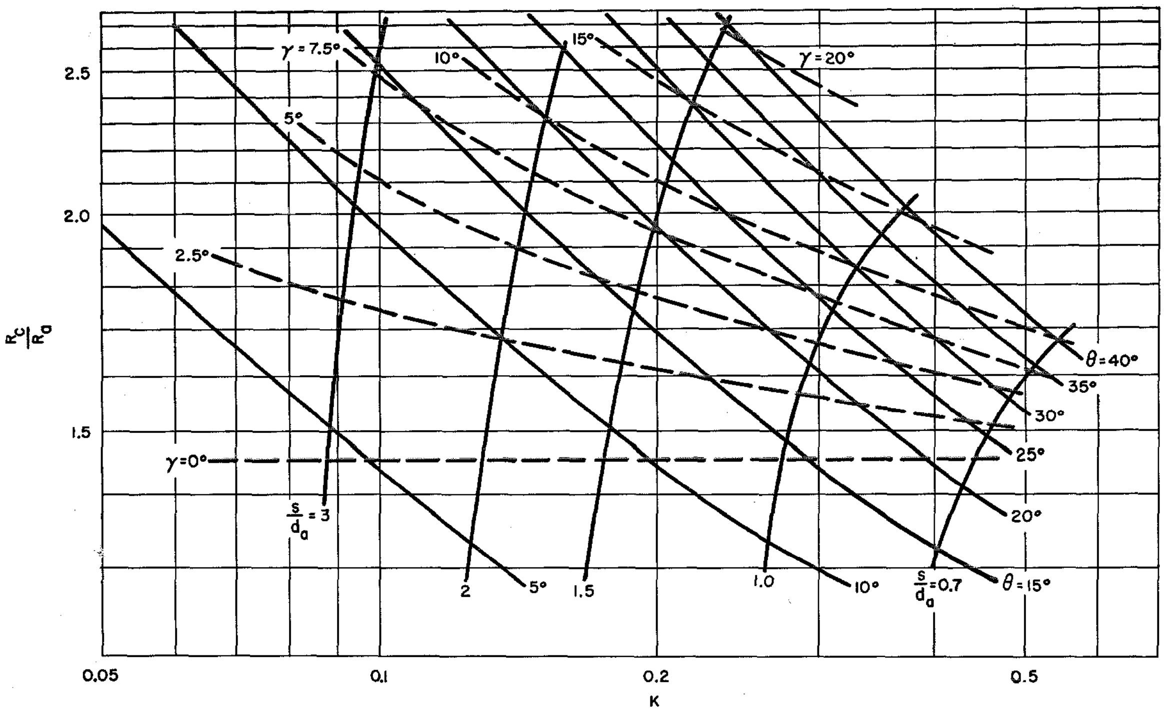

Fig. 2—Universal design chart, showing relationship of the four variables, K, R_c/R_a, θ, and γ.

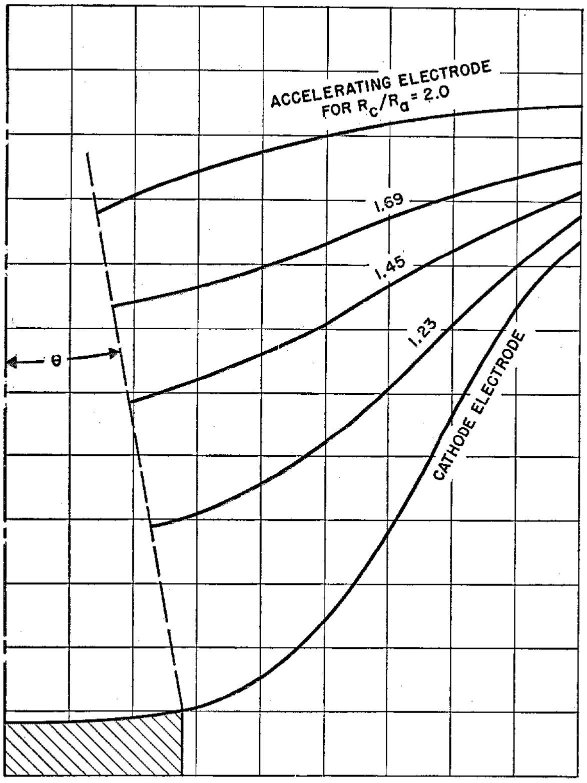

Fig. 4—Shape and placement of accelerating electrode for indicated values of R_c/R_a, where $\theta = 10$ degrees.

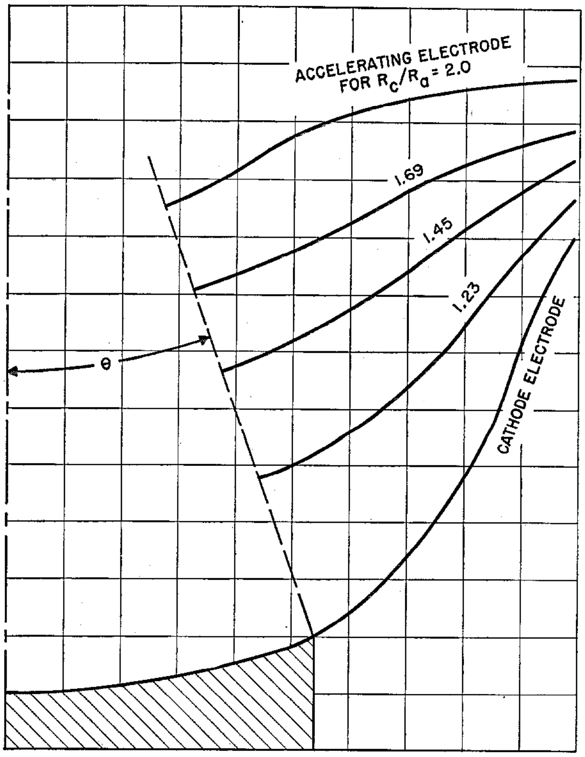

Fig. 5—Shape and placement of accelerating electrode for indicated values of R_c/R_a, where $\theta = 20$ degrees.

Fig. 2 (with reference to the dimensions illustrated in Fig. 1):

V = Beam voltage, kilovolts.
I = Beam current, amperes.
R_c = Cathode radius, spherical coordinates.
R_a = Aperture radius, spherical coordinates.
r_c = Cathode radius, cylindrical coordinates.
r_a = Aperture radius, cylindrical coordinates.
d_a = Aperture diameter, cylindrical coordinates.
s = Cathode-to-aperture spacing on the axis.
θ = Semiangle of the beam between cathode and aperture.
γ = Semiangle of the beam leaving the aperture.

Fig. 2 is a universal chart showing the relationship between the four variables K $(=I^{\frac{1}{2}}/V^{\frac{3}{4}})$, R_c/R_a, θ, and γ. Each point on the chart corresponds to a particular design of cathode structure. If any two variables are prescribed, ordinarily K and γ, then a point of the chart is determined and the corresponding values of the other two variables may be obtained. The electrode shapes for any cathode structure can be determined by methods described later in this paper. The electrode shapes for a wide range of cathode structures are given in Figs. 3 to 6.

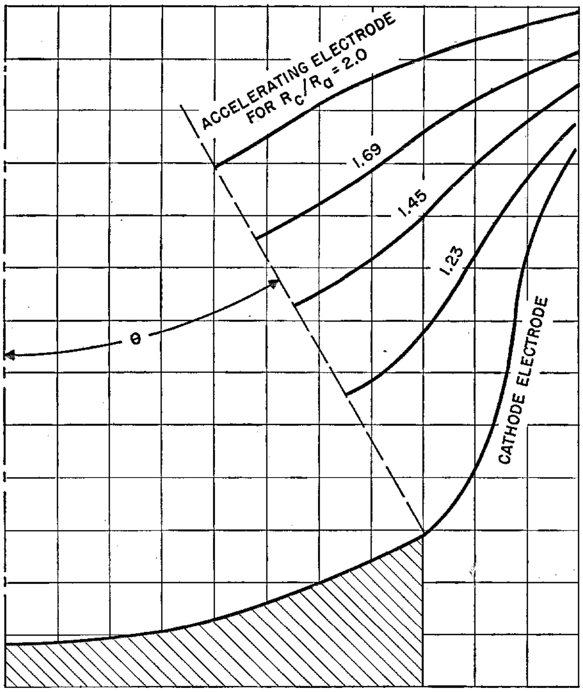

Fig. 6—Shape and placement of accelerating electrode for indicated values of R_c/R_a, where $\theta = 30$ degrees.

The relation between the four variables (Fig. 2) was determined as follows:

$$K=\frac{I^{\frac{1}{2}}}{V^{\frac{3}{4}}}=\frac{0.963\sin(\theta/2)}{\alpha}, \tag{1}$$

where α is the spherical space-charge-limited current flow factor:

$$\alpha=\log\frac{R_a}{R_c}-0.3\left(\log\frac{R_a}{R_c}\right)^2+0.075\left(\log\frac{R_a}{R_c}\right)^3+\ldots. \tag{2}$$

Values of α^2 as a function of R_a/R_c are given in Table I. The factor sin $(\theta/2)$ takes into account the fact that the beam is a cone of semiangle θ cut from the current flow between concentric spheres. From this relation, the solid curvilinear coordinates of θ were plotted as functions of K and the ratio R_c/R_a.

For the case of an aperture without a grid, consideration must be given to the effect of the distortion of the electric field caused by the aperture. The aperture acts as a divergent electron lens of focal length approximately $4V/E$, where V is the voltage between cathode and accelerating electrode, and E is the electric field for the space-charge-limited current flow on the cathode side of the aperture. It is assumed that the beam passes into field-free space beyond the aperture, and that the position of the lens is at the intersection of the axis with the spherical anode radius[4] R_a. The distortion of the spherical equipotentials by the aperture causing the lens action is shown in Fig. 7. The above assumptions yield the strongest possible lens action at the aperture. For an actual structure, the lens action will be somewhat weaker, and the beam on leaving the aperture will be slightly more convergent than the curves for γ (which were plotted from these considerations) indicate. The amount of lens action is independent of applied voltage (provided the cathode has sufficient emission for the current to follow the Child's law relationship) so that the shape of the beam is theoretically independent of cathode-to-accelerating-electrode voltage.

Once the values of R_c/R_a and θ have been determined, there remains the problem of determining the proper shape for the cathode and accelerating electrodes. Remembering that the electron beam is a cone cut from the inward radial current flow between concentric spheres, the electrodes must produce at the edge of the beam the same electric field as the rest of the spherical flow that has been cut away. To obtain the desired beam shape, the electric field just inside the beam (which conforms to the spherical space-charge-limited current flow relationships) must be matched by the electric field just outside the beam, which is the field produced by the properly shaped cathode and accelerating electrodes in charge-free space. The problem is that of shaping the electrodes properly to produce a field at the edge of the beam that has zero component perpendicular to the beam, and has a variation of potential with radius proportional to the factor $\alpha^{4/3}$.

TABLE I

α^2 AS A FUNCTION OF RADIUS

R_a/R_c	$(+\alpha)^2$	R_a/R_c	$(+\alpha)^2$
1.00	0.0000	6.5	13.35
1.05	0.0024	7.0	15.35
1.10	0.0096	7.5	17.44
1.15	0.0213	8.0	19.62
1.20	0.0372	8.5	21.89
1.25	0.0571	9.0	24.25
1.30	0.0809	9.5	26.68
1.35	0.1084	10	29.19
1.40	0.1396	12	39.98
1.45	0.1740	14	51.86
1.5	0.2118	16	64.74
1.6	0.2968	18	78.56
1.7	0.394	20	93.24
1.8	0.502	30	178.2
1.9	0.621	40	279.6
2.0	0.750	50	395.3
2.1	0.888	60	523.6
2.2	1.036	70	663.3
2.3	1.193	80	813.7
2.4	1.358	90	974.1
2.5	1.531	100	1 144
2.6	1.712	120	1 509
2.7	1.901	140	1 907
2.8	2.098	160	2 333
2.9	2.302	180	2 790
3.0	2.512	200	3 270
3.2	2.954	250	4 582
3.4	3.421	300	6 031
3.6	3.913	350	7 610
3.8	4.429	400	9 303
4.0	4.968	500	13 015
4.2	5.528	600	—
4.4	6.109	800	—
4.6	6.712	1 000	—
4.8	7.334	1 500	—
5.0	7.976	2 000	—
5.2	8.636	5 000	—
5.4	9.315	10 000	—
5.6	10.01	30 000	—
5.8	10.73	100 000	—
6.0	11.46	—	—

Taken from reference 2, p. 53.

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[4] From the geometry of Fig. 1 and the thin-lens formula, $\dfrac{\sin\theta}{\sin\gamma}=\dfrac{1}{1-\dfrac{R_a}{f}}$, where $f=4V/E$.

The electrode shapes can be determined most conveniently by use of the electrolytic tank.[3] An insulator is placed to correspond to the edge of the beam, so that the direction of the electric field must be parallel to the beam at its edge. Electrode shapes that will give the proper variation of potential along the insulator must then be determined. The analytical solution for a parallel beam of infinite width gives for the cathode electrode a plane at 67.5 degrees to the edge of the beam. For structures producing convergent beams, the conditions at the edge of the cathode, where curvature can be neglected, are the same as those for a parallel beam. The cathode electrode for all structures must therefore leave the cathode proper at an angle of 67.5 degrees to the edge of the beam, though this angle will hold only for a short radial distance. To simplify determining the shape of the remainder of the cathode electrode and the accelerating electrode, it was found convenient to set up vacuum-tube voltmeters along the insulator that represented the edge of the beam. The bias on these voltmeters was adjusted so that each meter would indicate half-scale with the desired potential at its probe. Copper strips representing the electrodes were then bent until all meters indicated half-scale. Multiple voltmeters were quite necessary, for without them it would have been very difficult to bend the strips to give the correct voltage at one point along the insulator without making the voltage at another point farther in error. Since the structures have cylindrical symmetry, the tank was tilted at angle θ to give a thin wedge of electrolyte tapering to zero thickness on the axis. A circuit diagram for the tank is shown in Fig. 8.

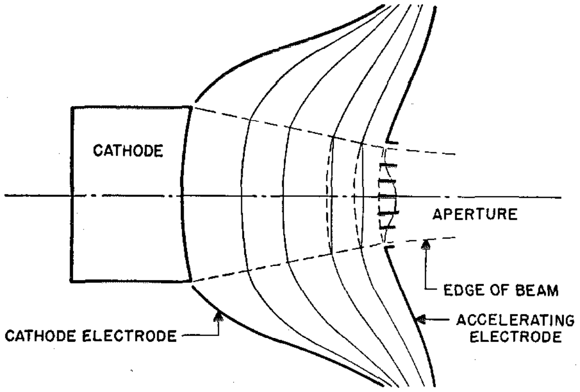

Fig. 7—Distortion of spherical equipotentials by the aperature with the resulting lens action. The dotted line indicates the correct equipotentials, the light solid line indicates the distorted equipotentials, and the small bars in the aperature indicate grid bars which may be used to maintain spherical equipotentials.

A series of electrode shapes for various angles of beam convergence θ are shown in Figs. 3 to 6. For each value of θ there is a family of accelerating-electrode shapes corresponding to particular values of the ratio R_c/R_a. *While for any given geometry there is theoretically only one correct shape for both electrodes, there are actually a large number of shapes departing slightly from the correct shape that will produce very nearly the correct field at the edge of the beam.* With the use of the electrolytic tank, "edge effects" for short electrodes are automatically taken into account, and the electrode shapes determined are silghtly distorted to compensate for them. It is also very

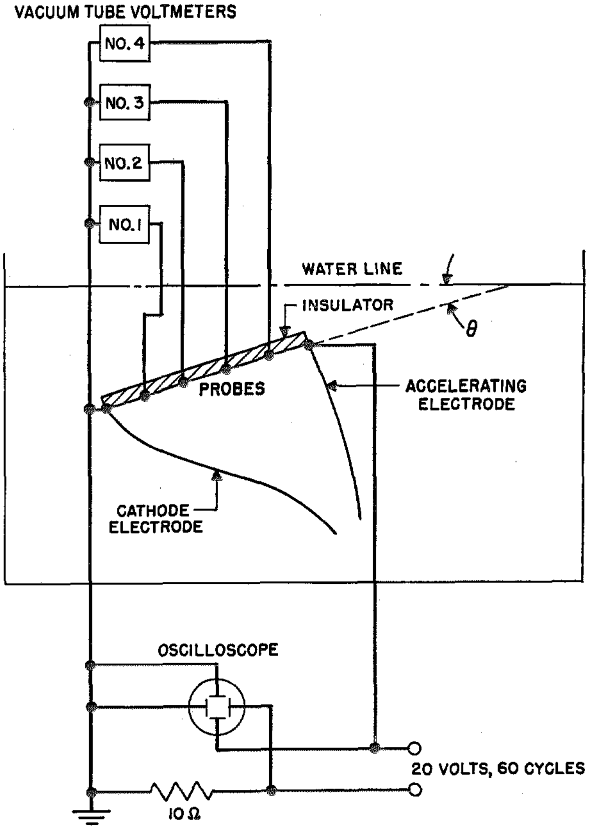

Fig. 8—Schematic diagram and vertical cross-section of electrolytic-tank circuit set up for determination of proper electrode shapes.

easy to determine the shortest electrodes that will produce the correct field at the edge of the beam. One electrode may depart slightly from the correct shape, and a corresponding shape found for the other electrode that will give the correct field.

Fig. 9 shows an experimental structure used to test the theory. The angle of convergence γ of the beam on leaving the aperture was not measured directly. Instead, the diameter of the beam at two measuring screens was observed. See Table II. A small axial voltage gradient was maintained along the entire length of the beam to remove any positive ions, since the calculation of beam diameter at the screens assumed full high-vacuum beam spread[5].

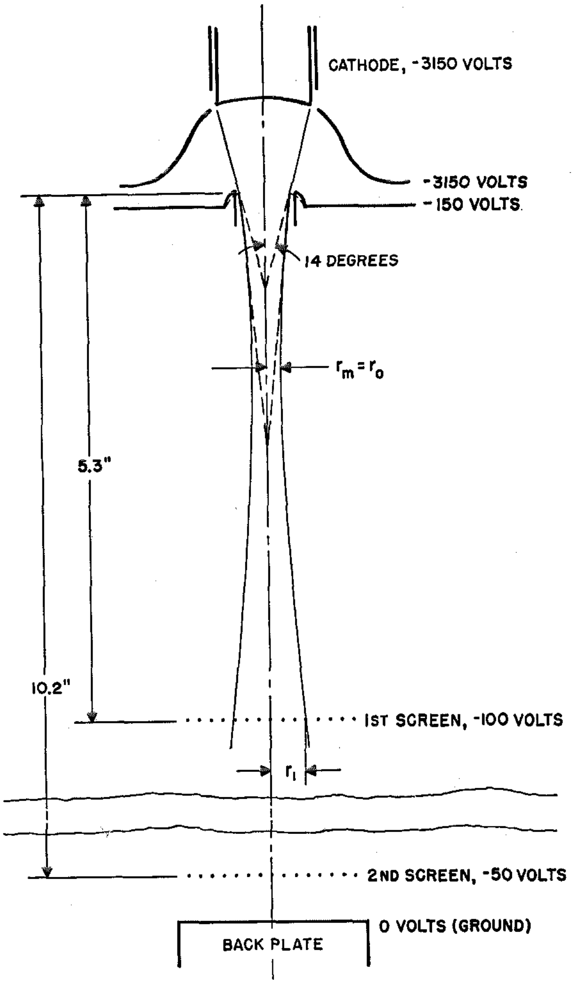

Fig. 9—Experimental structure used to test theory.

With the trend toward increasing the perveance of vacuum tubes, it is interesting to note the limitations on the maximum perveance that can be obtained from these structures. High-perveance structures lie at the extreme right of Fig. 2. In moving to the right along lines of constant θ, lower values of R_c/R_a are obtained. For a given value of R_c, the accelerating electrode must be brought nearer to the cathode to reduce the ratio R_c/R_a.

TABLE II

Calculated and Experimental Determination of Beam Convergence

	Voltage	Current in Amperes	Percent Current Through Aperture	Beam Diameter in Inches	
				1st Screen	2nd Screen
Calculated	1000	0.017	100	0.88	1.76
Experimental	1000	0.017	98.1	0.8	1.8
Calculated	2000	0.048	100	0.88	1.76
Experimental	2000	0.049	98.5	0.8	1.8
Calculated	3000	0.088	100	0.88	1.76
Experimental	3000	0.082	98.5	0.8	1.8

The following conditions apply: $K=0.13$, $R_c/R_a=2.06$, $\theta=14.3$ degrees, $\gamma=6.2$ degrees.

For the case of structures without grids, we must move to the right along lines of constant γ. This requires increasing the value of θ as well as reducing the ratio R_c/R_a. The divergent-lens action of the aperture becomes stronger as the accelerating electrode approaches the cathode, so the beam must start at a greater semiangle θ to have the same semiangle γ as it leaves the aperture. The limit is reached when the ratio of cathode–aperture spacing s to aperture diameter d_a becomes so small that the distortion of the field caused by the aperture (Fig. 7) is large at the cathode surface. When this occurs the current drawn from the center of the cathode drops below that drawn from its edges; the beam becomes nonuniform in current density, and the current becomes less than Fig. 2 indicates. A rough idea of the magnitude of this effect can be obtained from the solution for the axial electric field for the two-aperture electron lens.[6] The conditions in the lens only approximate those in our cathode structure, but the lens is considered because of its simple analytical solution. This solution gives a reduction in electric field at the center of the cathode of approximately 5 percent for $s/d_a = 0.70$, and 10 percent for $s/d_a = 0.56$. For convenience, contours of s/d_a are plotted in Fig. 2. The curves for θ and γ were arbitrarily stopped at $s/d_a = 0.70$. It should be remembered that this limitation is imposed by the current that can be drawn at a given voltage, and is not a limitation on the cathode-current density or the beam-current density. The practical limit of perveance that can be obtained from gridless structures is a subject requiring further experimental investigation.

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[5] B. J. Thompson and L. B. Headrick, "Space-Charge Limitations on the Focus of Electron Beams," *Proceedings of the I.R.E.*, v. 28, pp. 318–324; July, 1940.

[6] S. Bertram, "Determination of the Axial Potential Distribution in Axially Symmetric Electrostatic Fields," *Proceedings of the I.R.E.*, v. 28, pp. 418–420; September, 1940.

This design procedure is an extension of the work of J. R. Pierce. By means of the charts, it is possible to obtain immediately a cathode structure for any current and voltage and at any angle of convergence over a wide range of these variables. The experimental work was not extensive enough to insure that the extremely high transmission found in the sample structures tested would be obtained in all cases, but it is believed that beam transmissions of at least 85 percent should result from the application of these design charts. The maximum perveance that can be obtained from these structures is another subject requiring further experimental investigation.

Control of Electron-Beam Dispersion at High Vacuum by Ions

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Editor's Note: The results of a research project carried on at Stanford University during 1941 and 1942 under the sponsorship of the International Telephone and Telegraph Corporation were published in part in a book entitled The Production and Control of Electron Beams *by Karl Spangenberg, Lester M. Field, and Robert Helm. This limited edition was designated as "Secret" during the war, but is now declassified. This paper is based on parts of the book, which will not be reprinted in its entirety.*

THE POSSIBILITY that a commonly accepted theory of electron-beam action in a field-free space failed to account for certain dispersion effects led to the construction of apparatus to establish and to measure any discrepancy which might exist. The apparatus showed that such a discrepancy did actually exist and permitted quantitative measurements of the effect.

Briefly, the effect found was a dispersion of a high-density electron beam when passing through a field-free drift space at a gas pressure at which no dispersion was to be expected. It had previously been believed that such a beam would produce positive ions from the gas in the tube even at extremely high vacuum and, because of the absence of any external electric field in a drift tube, such ions would remain in the beam and effectively neutralize the negative space charge of the beam. The slow electrons produced in the ionization process would be removed from the beam and sent to the tube walls by the electric field produced by beam space charge before neutralization. This field would also cause the positive ions to move toward the beam axis. Consequently, such a beam would be expected to undergo no dispersion after reaching equilibrium, at least until a vacuum of the order of 10^{-10} millimeter of mercury was reached, at which pressure an insufficient number of gas molecules would exist for neutralization to occur. Unfortunately for such predictions, which were relied on in some commercial tube designs, the beam was found to begin its dispersion at a pressure of 10^{-6} millimeter, or at a pressure 10 000 times higher than expected.

Fig. 1 shows the manner in which the amount of dispersion undergone by a beam was found to vary for different velocities of the electrons. Low-velocity (or low-voltage) beams spread appreciably at a much higher vacuum than did high-voltage beams, and also increased their spread at a lower rate as the pressure dropped, although they ultimately reached much greater total dispersion. An optimum voltage was found to exist for which beam dispersion began at the poorest vacuum. Higher- and lower-voltage beams began their dispersion at a better vacuum.

A search for an explanation led to an approximate theory of ion removal which has had considerable success in predicting all the effects observed. It has also led to the invention of means for preventing such effects. The theory quantitatively predicts the amount of beam dispersion and the pressures at which such dispersion begins to be appreciable; it uses only standard physical constants. It also predicts the optimum voltage at which the effects begin and seems to check closely all of the observed phenomena.

Recombination, wall charges, increased gas temperature, and ion velocities achieved by collision during ionization, have been ruled out as possible explanations by consideration of their magnitudes and effects.

1. *Ion-Removal Process*

The proposed theory is based on the idea of an ion sink, or in other words, an external field which reaches a small distance into the field-free space and continuously removes ions from this portion of the beam. Such a field is known to exist at the aperture through which the electron beam first enters the field-free space, or drift tube; this field is an extension of the strong electron-accelerating field in the near vicinity. The important question is to ascertain how far down the beam this ion removal is effective. Since removal of positive ions at one section of the beam (by an external field, for example) causes the

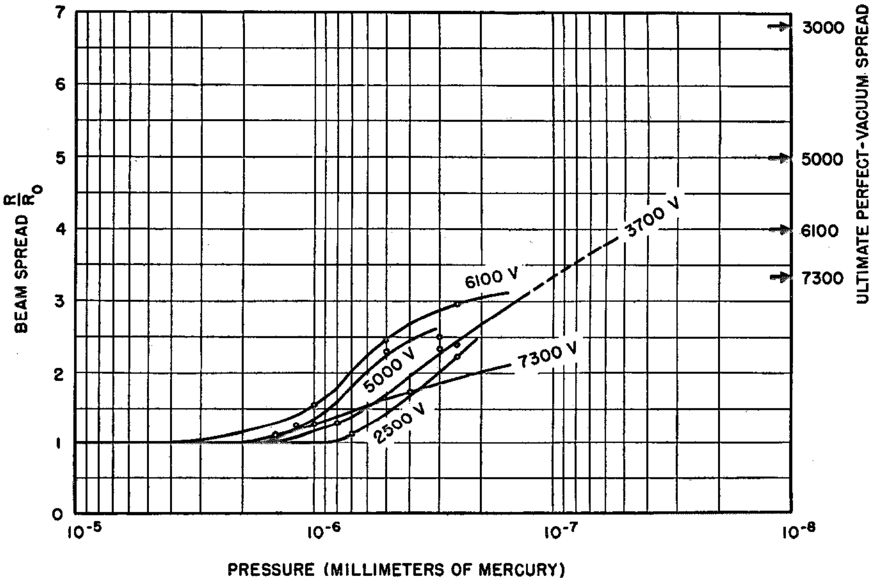

Fig. 1—Results of beam-spread measurements. Beam current = 50 milliamperes. Drift-tube length = 36 centimeters. Beam-current density = 0.1 ampere per square centimeter.

center of the beam at this section to become more negative in potential, ions from nearby sections tend to flow toward this point. This provides a mechanism whereby ions from points still farther down the beam start flowing toward the sink. Thus ion removal from much greater distances than the actual field penetration is possible. It remains now to formulate mathematically the process involved, to determine whether the effect is negligible or important, and to find in what manner it depends on the field at the aperture, pressure (which influences the rate of production of ions), beam voltage, and beam current.

Consider now Fig. 2, in which is depicted a drift tube, or field-free space (the effect propagating back to the sink occurs in field-free space away from the aperture). The electrons enter at the left with velocity v and traverse the space at essentially this velocity until they hit the back plate at $X = d$. Throughout the volume traversed, they produce ions at a constant rate of n per centimeter. Then the number of ions per second produced in a length of beam dX will be ndX.

Under the equilibrium conditions, these ions and all other ions produced in the beam must flow to the left under the influence of potentials produced by space charge only, and they must be removed from the beam at the ion sink at the same rate as that at which they are being produced. Then, some equilibrium number of ions may be considered to exist at each part of the beam, depending on the number and velocity of ions passing that point. However, the ions passing any given point have a range of velocities determined by the potential they have fallen through which, of course, depends on the distance between the starting point and the point in question. Further, the potential at each point depends on the above result, and consequently an integral relation for the total effect may be written.

Consider conditions at the plane X_0, through which all ions produced to the right of this plane are flowing (Fig. 2). As a first approximation, consider the ions to be flowing principally on the axis, since the form of potential variation across the beam produces an inward radial force on the ions except where the electron space charge is completely neutralized, and consequently all ions in the unneutralized regions tend to move toward or oscillate closely about the axis. The ions will be moving through the plane X_0 with velocities dependent on $(V_X - V_{X_0})^{\frac{1}{2}}$, where V_X is the potential at the center of the beam at plane X, where the ions are produced, and V_{X_0} is the potential at the center of the beam at plane X_0. At X_0 the current of ions, or number of ions per second, from the section dX will be ndX, as described before. These particular ions have a

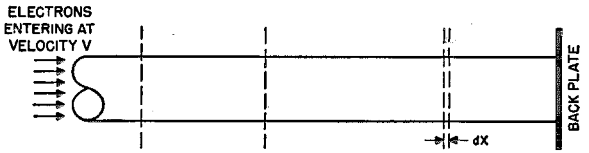

Fig. 2—Electron beam in field-free space.

velocity $k(V_X - V_{X_0})^{\frac{1}{2}}$, so the density at X_0 of ions from section dX at X only will be

$$d\rho_{X_0} = \frac{n\,dX}{k(V_X - V_{X_0})^{\frac{1}{2}}}, \text{ ions per centimeter.} \qquad (1)$$

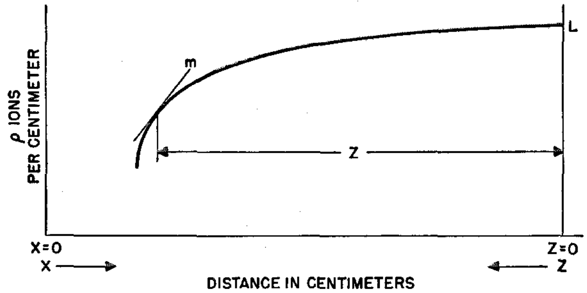

Fig. 3—Universal ion-density curve.

The total ion density at X_0 will be given by

$$\rho_{X_0} = \int_{X_0}^{d} \frac{n\,dX}{k(V_X - V_{X_0})^{\frac{1}{2}}}, \text{ ions per centimeter.} \qquad (2)$$

Now an application of Gauss's law shows that in an electron beam of constant density $\rho-$, on which is superimposed an ion beam of varying density $\rho+$, the potential difference between points on the center of the beam at, for example, positions X and X_0 is exactly equal to the difference in positive-ion charge density at the two points. Or

$$V_X - V_{X_0} = \rho_X - \rho_{X_0}, \qquad (3)$$

where V is in electrostatic volts and ρ is in electrostatic units per centimeter.

It should be noted that here and elsewhere throughout this paper charge densities are given per linear dimension, and all effects are independent of the area of the beam involved except where otherwise noted.

Now, if in the integral formulation of the problem, $\rho_X - \rho_{X_0}$ is substituted for $V_X - V_{X_0}$, the equation is an integral equation from which the manner in which ρ varies with X can be found. It is now

$$\rho_{X_0} = c\int_{X_0}^{d} \frac{dX}{(\rho_X - \rho_{X_0})^{\frac{1}{2}}}, \qquad (4)$$

where

$$c = n/K. \qquad (5)$$

We seek a solution for ρ as a function of X, $\rho(X)$, which will satisfy this integral for various values of the constants.

The solution of this integral equation (see 6.1) is as follows:

$$3\pi c(d-X) = 2(L+2\rho)(L-\rho)^{\frac{1}{2}}, \qquad (6)$$

where L is the value of ρ at $X=d$, and c is a constant depending on the rate of production of ions, their mass, and their charge.

It is more convenient to use $Z=d-X$ as the coordinate for distance along the beam, where Z is now the distance back from the point of maximum ion density. This is usually at the back plate or collector, although it may be at any point along the beam at which ions reach a limiting density and beyond which they are removed by another process (for example, normal radial flow to the drift-tube walls under equilibrium conditions in an electron beam). Then,

$$3\pi cZ = 2(L+2\rho)(L-\rho)^{\frac{1}{2}}. \qquad (7)$$

This equation defines a form of variation, which can be represented by a curve, and which must be followed by ion density under all conditions for which this process occurs. The relative scales of this universal curve depend, however, on the value of the constant c (a function of pressure, current, and voltage), and the value of L. Fig. 3 shows the form of this universal curve. Another set of relations, more useful for calculation, gives the same result in terms of c and

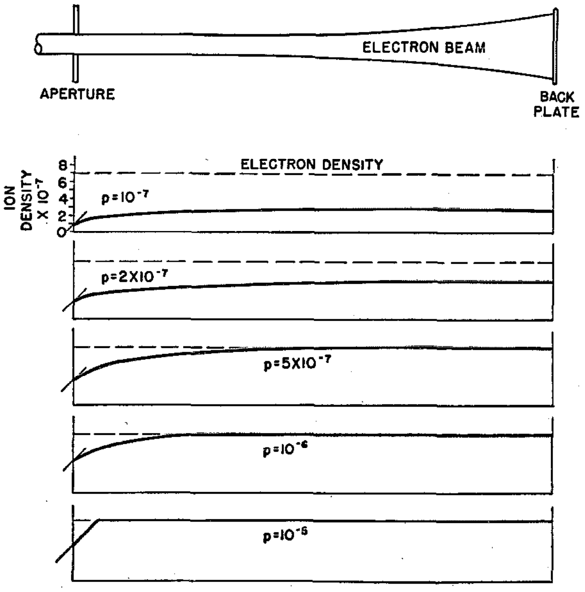

Fig. 4—Variation of ion density with pressure *p* in millimeters of mercury. The dashed lines plot the electron density.

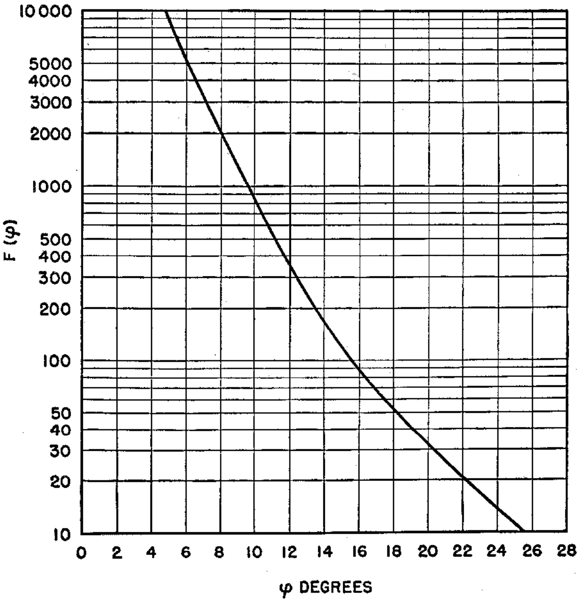

Fig. 5—Plot of the function $F(\varphi)$.

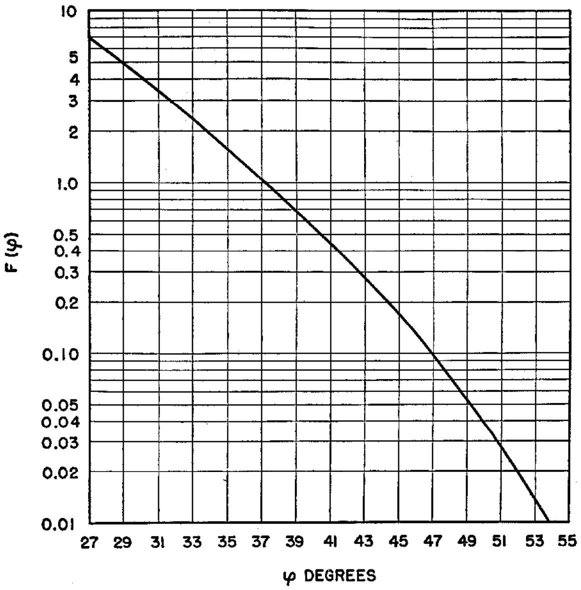

Fig. 6—Plot of the function $F(\varphi)$.

gradient m at a specified point along the beam (at the aperture).

It should be noted that the gradient becomes infinite at a point on the curve. This merely means that an infinite gradient could extend the process from the point in question forward in the manner shown. Since a finite gradient exists at the aperture where the process starts, the actual ion density varies, according to the curve, from the point of tangency of the line with slope m (m = gradient at aperture) up to a maximum value L. Infinite gradient is now no longer necessary because to the left of the point of tangency a new process, i.e., an external field, is removing the ions.

If, starting with some definite value of gradient, the ion-density curve rises so fast that it reaches a value equal to that of the electron density of the beam before reaching the back plate, the ion density will be postulated to rise no higher, but rather remain constant and equal to the electron density for the remaining distance to the back plate. The excess number of ions formed in this region will be removed from the beam radially to the drift-tube walls, for their presence in excess of the electron density produces a field which acts radially outward on positive ions. The manner in which the universal ion curve of Fig. 3 adjusts itself to varying pressure at a constant aperture gradient (determined by external factors) is shown in Fig. 4 for a typical beam (an approximate representation of the 6000-volt-beam condition). This curve shows the various conditions just described.

When the constants are evaluated (see 6.3), the fundamental equation becomes

$$\frac{pi}{V}Z\cdot 1.34\cdot 10^{22}=(L+2\rho)(L-\rho)^{\frac{1}{2}}, \tag{8}$$

where

p = pressure in millimeters of mercury,
i = current in amperes,
Z = distance in centimeters from the back plate, or positions of L to the position at which ρ is the ion density,
V = beam voltage in volts,
L = maximum linear ion density in ions per centimeter and,
ρ = linear ion density in ions per centimeter at any position Z.

The equation as given above cannot be readily solved directly for L and ρ when we know p, i, V, and Z (even though we know $\rho = L$ at $Z = 0$), and also know $d\rho/dZ$ at one point.

A solution in parametric form has been worked out which gives the answers directly in terms of known constants. In terms of a parameter $F(\varphi)$, for which curves are given in Figs. 5 and 6, the solution in practical units (see 6.2) is

$$\frac{p^2 i^2}{d V^2 m^3}\cdot 7.96\cdot 10^{22}=F(\varphi). \tag{9}$$

Knowing $F(\varphi)$, the curves give the value φ, and then

$$L=4.47\cdot 10^{14}\left[\frac{pid}{V\sin\left(\frac{3\varphi}{2}\right)}\right]^{\frac{2}{3}}, \tag{10}$$

where

p = pressure in millimeters of mercury,
i = beam current in amperes,
d = distance from aperture to point of maximum ion density (where L occurs),
V = voltage corresponding to beam velocity,
m = gradient at the aperture in volts per centimeter,
L = maximum ion density in ions per centimeter, and

$$F(\varphi)=\frac{\cos^3\left(\frac{3\varphi}{2}\right)}{\sin\left(\frac{3\varphi}{2}\right)\sin^3\varphi}. \tag{11}$$

(Curves are given of this function.)

Thus we determine L in terms of known conditions, and therefore know the degree of neutralization of the electron beam.

2. *Comparison of Theoretical and Measured Results*

Values calculated from the theory proposed in Section 1 may now be compared with the measured values of beam spread versus pressure given in Fig. 1.

For this purpose, the values 3000, 6000, and 7300 volts, all at 0.050 ampere total current, have been selected for calculation. The results are given in this section and are compared with the observed beam-spread values.

In addition to the voltage, current, and length of beam, the gradient at the aperture must be known. This is the gradient produced by field penetration through the aperture from the main accelerating field into the previously field-free drift space.

For the aperture-cylinder configuration used, the field penetration would be approximately 0.1 percent of the maximum field present in the accelerating region; field plots show that the field penetrating down the cylinder would be approximately 1 percent of the maximum field and this is reduced by a factor of 10 in passing through the aperture.

The structure used was of such form that the main accelerating field in volts per centimeter was fairly closely $V/16.6$ (from a field plot of the configuration used), so that the maximum accelerating fields came out as indicated in Table I.

TABLE I
MAXIMUM ACCELERATING FIELDS

V in Volts	E (Maximum) in Volts per Centimeter	m or E Through Aperture in Volts per Centimeter
3000	180	0.180
6000	360	0.360
7300	438	0.438

The values given above are used in the calculations. The results are not particularly sensitive to a variation in field however, and so slight errors in this estimate are not serious.

The detailed calculation of theoretical beam spread for the conditions specified, where $V=3000$ volts, follows from these preliminary calculations:

$$\left.\begin{aligned}F(\varphi)&=7.96\cdot 10^{22}\frac{i^2p^2}{dV^2m^3}\\&=\frac{7.96\cdot 10^{22}\times 0.050^2\times p^2}{36\times 3000^2\times 0.18^3}=1.05\cdot 10^{14}\times p^2.\end{aligned}\right\} \tag{12}$$

Then

$$L=4.47\cdot 10^{14}\left[\frac{pid}{V\sin\left(\frac{3\varphi}{2}\right)}\right]^{\frac{2}{3}},$$

or, if we let

$$\xi=\left[\frac{1}{\sin\left(\frac{3\varphi}{2}\right)}\right]^{\frac{2}{3}},$$

then

$$L=3.2\cdot 10^{12}\times p^{\frac{2}{3}}\times\xi, \tag{13}$$

when $d=36$ centimeters.

For the case where L exceeds the electron density, the length of beam over which the ion density is less than the electron density is of interest. As the electron density is $9.6\cdot 10^7$ electrons per centimeter, the value of d (distance from the aperture) at which the ion density equals the electron density is, from (10) with $9.6\cdot 10^7$ for the value of L,

$$d^{\frac{2}{3}}=\frac{3.3\cdot 10^{-4}}{p^{\frac{2}{3}}\xi}, \tag{14}$$

from which d may be found.

It is of interest to know the value to which L rises at the end of the beam (36 centimeters from the aperture), if this value is less than the electron density. Then the beam is obviously less than fully neutralized, and the value of L helps in estimating the amount by which the beam will spread. The value of L at the end of the beam is given by (13).

If L will rise to greater than the electron density in 36 centimeters, the value of L is no longer of interest, but instead the value of d at which the ion density first reaches the electron density becomes important and is given in (14).

The two conditions occur in the examples of Fig. 4, which show the ion density variation from highest to poorest vacuum. The resulting beam spread may then be estimated with the aid of the universal-beam-spread curve of Fig. 7, which is based on ion-free or perfect-vacuum conditions. An exact solution of the beam spread when the charge density varies with length in the fashion shown in Fig. 4 would probably be very difficult, although it makes an interesting problem. For the purposes of this calculation, such an added complication is hardly justified, so a standard method of estimating the spread is used. This means that the precise values of spread at conditions between no spread and perfect-vacuum spread may be somewhat in error, but relative values for different pressures and for different voltages may have some meaning because the same method of estimation is used in each case.

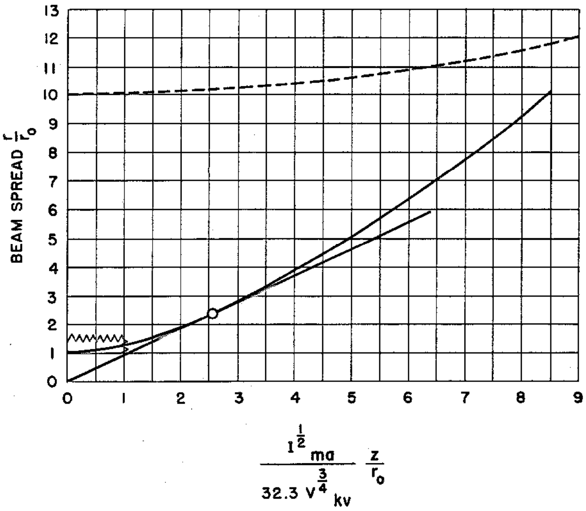

Fig. 7—Universal beam-spread curve (negative space-charge effect).

$$\frac{Z}{r_0} = \frac{V^{\frac{3}{4}}}{I^{\frac{1}{2}}} \cdot \frac{1}{2^{\frac{1}{4}}} \left(\frac{e}{m}\right)^{\frac{1}{4}} \int_1^R \frac{dR}{(l_n R)^{\frac{1}{2}}}, \text{ electrostatic units,}$$

$$R = \frac{r}{r_0}.$$

That part of the curve within the wavy line at the origin is plotted at 10 times size in the upper dashed curve. Both ordinate and abscissa scales should be divided by 10 in using this curve.

TABLE II

CHARACTERISTICS OF A 3000-VOLT BEAM

p in Millimeters of Mercury	$F(\varphi)$	φ degrees	ξ	L Ions per Centimeter	d in Centimeters	R/R_0 Beam Spread
$5\cdot10^{-6}$	260 000	2	7	—*	0.064	1.0
$1\cdot10^{-6}$	105	15.5	1.86	—*	2.35	1.1
$5\cdot10^{-7}$	26.0	21.0	1.54	$31\cdot10^{7}$*	6.30	1.5
$1\cdot10^{-7}$	1.05	36.9	1.14	$7.8\cdot10^{7}$	50†	3.8
$1\cdot10^{-8}$	0.0105	53.5	1.005	$1.5\cdot10^{7}$	—†	6.0
0	—	—	—	—	—	6.8

* Greater than electron density of $9.6\cdot10^{7}$.
† Longer than the beam length of 36 centimeters.

The method used was basically as follows: A beam just reaching complete neutralization at the back plate (corresponding to the middle diagram of Fig. 4) was considered to spread just *half* as much as a completely unneutralized beam, as the lack of neutralization in the initial portion of the travel would have a relatively great effect on this spread. *A beam reaching the full neutralized value at a fraction of the total length of the beam is considered to spread just half as much as an unneutralized beam would spread for this length, and then to continue to spread in a straight line at the angle reached until it gets to the back plate.* This implies complete neutralization in all the beam after the ion density equals the electron density. An excess number of ions over electrons presumably would not be found in the beam, as the radial field would repel them to the walls as previously described. Where the ion density fails to reach the electron density even at the back plate, the spread is taken as between half and all of the completely unneutralized spread by an amount proportional to the fractional neutralization at the back end. The results are then as given in Table II for the 3000-volt beam. The factor for use in Fig. 7 is $Z/5.84$.

Plots of similar calculations carried out for 6000 volts and 7300 volts, together with the results for 3000 volts, are presented in Fig. 8.

A direct comparison of these theoretical results with measured results of beam spread as shown in Fig. 1, may now be made. The following correlations may be observed to exist:

A. Experimentally, beams of all voltages at the specified current were found to start spreading at a pressure of approximately 10^{-6} millimeter of mercury. The theoretical curves also have this property.

B. The values of beam spread at any particular pressure check fairly well despite the approximate method of calculating this spread (described in the analysis of the 3000-volt beam).

C. In the measured results, the higher-voltage beams (up to a critical voltage) reached their limiting spread quickly as the pressure dropped, while lower-voltage beams increased their spread gradually with decreasing pressure. Above the critical voltage (approximately 6000 volts), spread again increased gradually with falling pressure. The theoretical curves show precisely these properties.

D. Perhaps the most unexpected correlation is the verification by the theory of a second-order effect which was just barely indicated on the measured results because of the difficulty of accurate measurement of small changes in beam spread. Measurements taken showed a small variation in the pressure at which dispersion began for various beam voltages. Dispersion began soonest (at poorest vacuum) for a beam of approximately 6000 volts. Higher- or lower-voltage beams began to disperse only at a better vacuum. The theoretical curves of Fig. 5 show that the theory predicts precisely such a variation, and indicates the optimum voltage (or worst voltage if the effect is considered as one to be avoided).

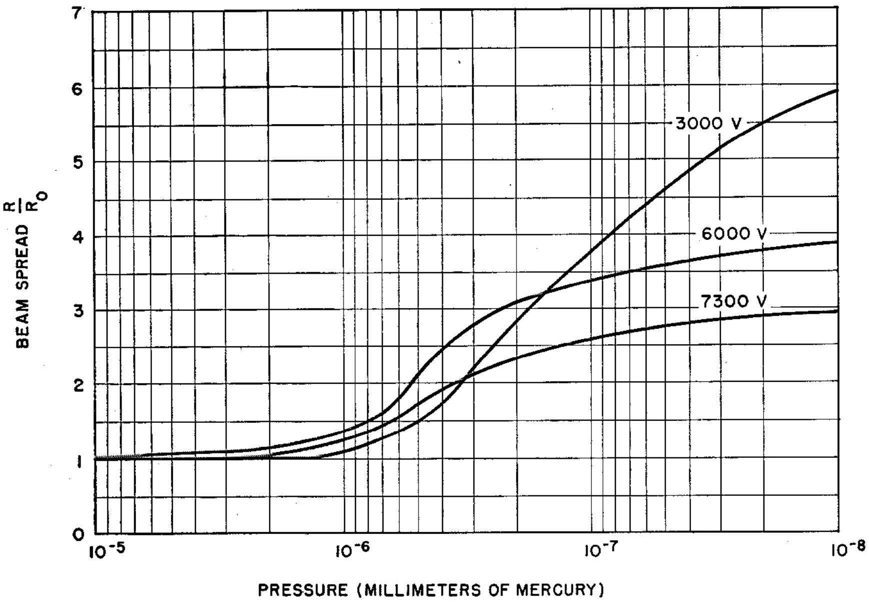

Fig. 8—Beam spread predicted from theoretical equations. Beam current = 50 milliamperes. Length = 36 centimeters. Beam-current density = 0.1 ampere per square centimeter. Compare with Fig. 1.

To summarize, the theory gives calculated values which agree closely with the measured values of the pressure at which the dispersion occurs, the amount of dispersion, and the variation in dispersion with voltage. In addition, a test of the basic premise of the theory is described in Section 4. The test appears to verify this premise and has led to a practical application of the theory to beam tubes.

3. *Apparatus and Experimental Procedure*

The apparatus for studying the effect of voltage, current, and pressure on the spread of a parallel beam consisted in part of a cathode and an accelerating electrode, followed by a focusing cylinder to form a parallel beam. There was then a limiting aperture, followed by some apertures in thin metal sheets placed so that the position at which the sides of the beam were parallel could be determined. This was followed by a long drift tube at the end of which was placed a thin metal target. The heating of this target gave an indication of the diameter of the beam. An ionization gage for measuring the pressure

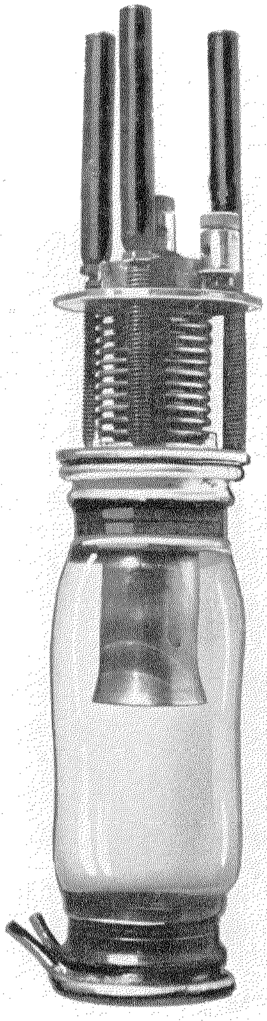

Fig. 9—Cathode structure.

was connected to the system near the middle of the drift space.

The cathode housing is shown in Fig. 9. The cathode diameter was about 1 inch, and it was moved inside of the surrounding cylinder by adjusting screws connected to a sylphon bellows.

The system was continuously pumped, and was held together with metal-to-glass seals and solder-filleted joints which had to be water cooled during operation.

The accelerating electrode was a cylinder supported between glass insulating sections. The focusing electrode was a copper cylinder connected to a heavy copper aperture as shown in Fig. 10. This picture also shows some 0.010-inch molybdenum disks with apertures which were used to test the parallel condition of the beam. This was indicated by the heating on the edges of the apertures; when the first and third apertures heated evenly and the middle aperture heated to a lesser degree, the beam was considered to be parallel near the middle aperture. This indicating method was later superseded by a system of screens of woven 0.001-inch tungsten wire which served the same function, but which was a much more sensitive indicating arrangement.

The first target used was a 0.010-inch disk of tantalum. However, the high currents and

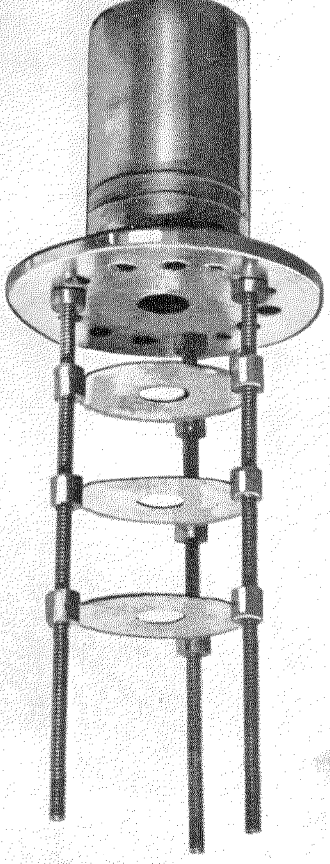

Fig. 10—Focusing electrode and molybdenum disks.

voltages used melted this target. The final arrangement consisted of a tungsten screen supported above a copper target which was cooled by circulating water (see Fig. 11).

The power-supply arrangement and connection diagram are shown in Fig. 12. The initial tungsten screens were viewed through the grill visible on the upper portion of the long copper drift tube shown in Fig. 11. This photograph shows the drift-tube assembly dismantled.

4. *Experimental Test of an Ion Trap*

The effect of field penetration through the aperture as described in Section I may be tested quite directly, and this test has been made.

Inasmuch as the process which removes ions depends on the gradient of potential at the aperture and calculations show that very minute gradients are sufficient to carry on the process, the most certain method of stopping the process would appear to be a reversal of the gradient at the aperture which would trap the ions in the drift space. A similar trap at the back plate would insure that ions formed in the beam would not flow out at either end and so could only leave the beam radially. Since this could happen only when the number of ions exceeds the number of electrons in the beam, the space charge of the beam would then be effectively neutralized.[1]

Apparatus was constructed and tested using a solid drift tube, insulated so that it could be maintained at a slight negative potential with respect to the aperture and back plate. The form of this equipment, potentials, and gradients involved are shown in Figs. 11 and 13. As a drift tube usually collects a small current of stray electrons, merely placing a resistor in an external circuit between the drift tube and the aperture can usually provide the necessary potential. Should stray-electron current prove in-

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[1] K. Spangenberg, L. M. Field, and R. Helm, U. S. Patent Application 447 194 (1942); and L. M. Field, U. S. Patent Application 601 095 (1945).

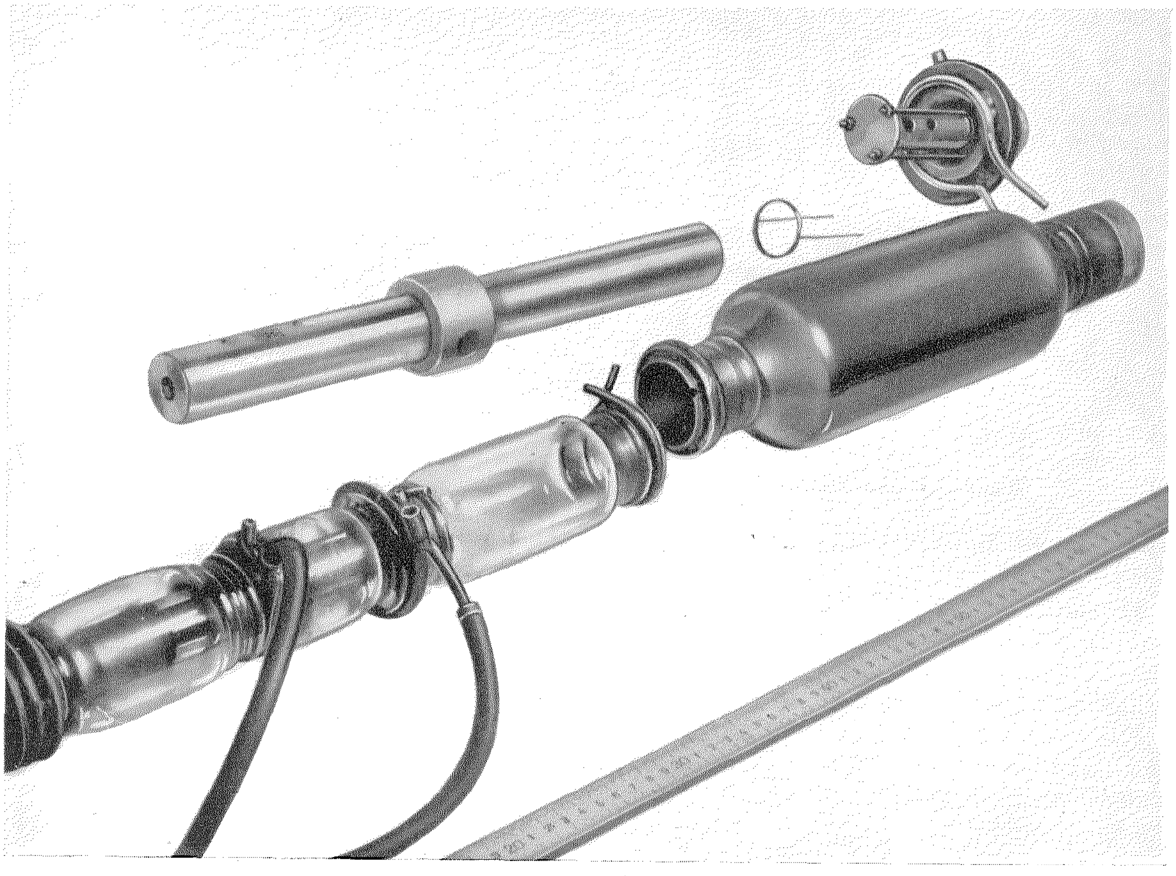

Fig. 11—Drift tube, target, and tungsten screens.

sufficient for this purpose, an auxiliary voltage supply could be used.

In a test made at a pressure of $5 \cdot 10^{-7}$ millimeter of mercury, a beam spreading to twice its original diameter in 36 centimeters at 7300 volts immediately stopped spreading when as little as 15 volts was applied between the drift tube and aperture in such direction as to reverse the gradient at this point. A 15-volt variation in 7300 volts would have no appreciable lens effect and this result may be considered as a verification of the importance of the aperture gradient in determining neutralization by positive ions.

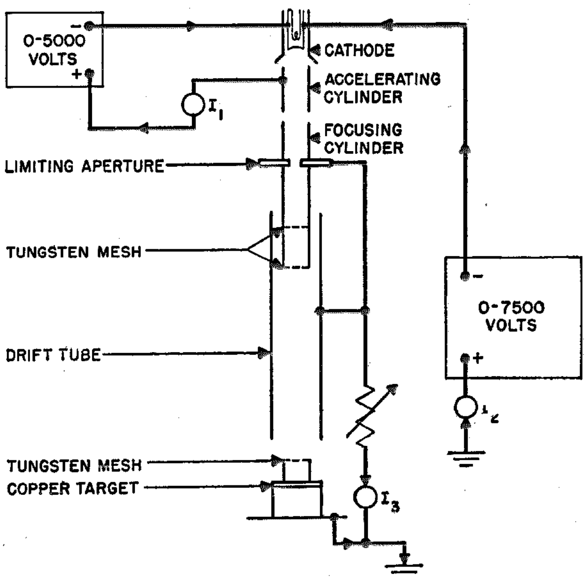

Fig. 12—Schematic diagram of apparatus.

Variations of beam spread with trapping voltage are given in Table III. As an auxiliary check, increasing the pressure to $5 \cdot 10^{-6}$ millimeter also served to stop the spread. Of course, increasing the pressure is impractical, for tubes are normally used as sealed-off devices; this merely demonstrated that the effect of an extremely good vacuum on ion-space-charge neutralization could be prevented by the use of an ion trap as suggested here.

TABLE III

Variation of Beam Spread with Ion-Trap Voltage

Pressure in Millimeters of Mercury	Voltages		Beam Current in Amperes	Drift Tube-to-Aperture Voltage	Spread R/R_0
	V_1	V_2			
$5 \cdot 10^{-7}$	1500	5800	0.130	0	2.0
$5 \cdot 10^{-7}$	1500	5800	0.130	7.5	1.2
$5 \cdot 10^{-7}$	1500	5800	0.130	15.0	1.1
$5 \cdot 10^{-7}$	1500	5800	0.130	30.0	1.0+
$1.1 \cdot 10^{-6}$	1500	5800	0.130	0	1.7
$3 \cdot 10^{-6}$	1500	5800	0.130	0	1.1
$5 \cdot 10^{-6}$	1500	5800	0.130	0	1.0+

It was noted in the test of ion-trap voltage just described that, for the particular beam used, 15 volts stopped practically all spreading and higher voltages produced very little increased effect. It is of interest to note that this voltage is only slightly greater than the difference in voltage between the edge and the center of the beam when it consists of unneutralized electrons only. This difference is approximately 10 volts for the beam tested.

The reason for the comparable values is evident when it is considered that on the cathode side of the aperture the ions are always removed so fast that the electron beam is essentially unneutralized. The center of the beam is then 10 volts negative with respect to the edge. Thus a negative gradient through which ions can flow will always be present until the drift-space potential is lowered by this 10 volts plus an additional voltage sufficient to reverse the gradient

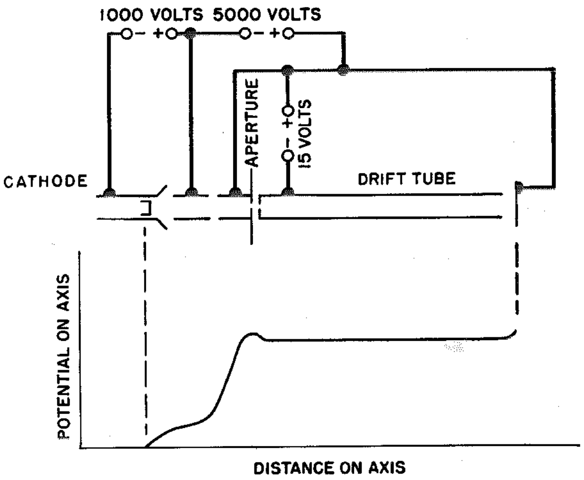

Fig. 13—Ion-trapping apparatus and potentials.

produced by field penetration. Fig. 14 illustrates the effect. Dotted lines show the potential at the center, and solid lines the potential at the edge of the beam.

Apparently the voltage difference required between the drift tube and its ends (back plate and aperture) to form an ion trap must exceed 10 volts for a 50-milliampere 6000-volt beam. The value of trapping voltage which must be exceeded to maintain ion neutralization varies directly with the charge density, and so would increase directly as the current in the beam and vary inversely as the square root of the voltage of the beam. For beams of much higher current, the ratio of the radius of the beam to the radius of the surrounding drift tube becomes of importance.[2]

The use of ion traps to keep beams from spreading at high vacuum greatly extends the possible limits of operation. Electron beams up to the limiting current permissible in the presence of ions,[3] which is $190 \cdot 10^{-6} V^{\frac{3}{2}}$ ampere, where V is in volts, presumably could be maintained by ion cores at even extremely good vacua (for example, 10^{-9} millimeter of mercury) if sufficient volume is provided in the tube. Because an ion once in the beam cannot leave it, and as virtually no recombination takes place, it appears that most of the molecules in a given tube at very high vacuum would soon be found as ions in the beam.

When pulsing of the beam is desired, the time required for the ion core to build up is of importance. This time is independent of the amount of current in the beam, but is directly proportional to the voltage of the beam and inversely proportional to pressure as

$$T = 3.6 \cdot 10^{-15} \frac{V}{p} \text{ seconds}, \qquad (15)$$

where T is the time necessary to build up an ion core to neutralize the electron beam fully, and V and p are in volts and millimeters of mercury, respectively. A typical value for a 1000-volt beam at $p = 10^{-7}$ is 36 microseconds. At $p = 10^{-6}$, this would be reduced to 3.6 microseconds, but this is probably too high a pressure for commercial use.

Therefore, to use pulsed beams, pulses of a duration greater than the times mentioned above must be used, or some auxiliary means must be found for maintaining the ions in position between pulses. One possible method of doing this is the use of a pilot beam, of much lower voltage and somewhat lower current than the main beam, to maintain the neutral condition in the drift tube. The current necessary would decrease as the square root of the pilot-beam voltage.

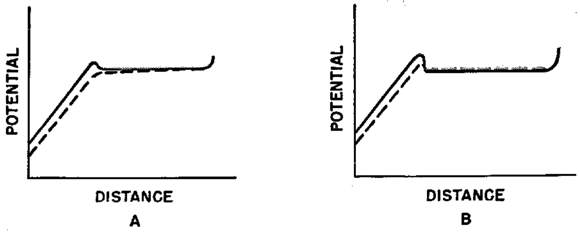

Fig. 14—Potentials at center and edge of beam during ion trapping.

It should be kept in mind that the use of the ion-trapping arrangement described here can be successful in reducing or removing space-charge repulsion effects in electron beams in field-free space only if every precaution is taken to prevent stray fields (from wall charges, insulators, nearby electrodes, or nearby conductors) from acting on the beam. Such a field causing ions to leave the beam at any point, or even tending to increase ion density in one region as compared with another, can result in the removal of virtually all ions from the beam through the affected region, with consequent failure of the ion-trapping action.

5. *Acknowledgment*

The authors wish to credit Mr. Paul Rosenbloom, formerly of the Department of Mathematics at Stanford University and now at Brown University, with the solution of equation (4) for ion flow, as presented in Section 6.1. They wish to acknowledge also his supplying of the essential method of the parametric solution given in Section 6.2.

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[2] D. P. R. Petrie, "Effect of Space Charge on the Potential and Paths of Electron Beams," *Electrical Communication*, v. 20, n. 2, pp. 100–111; 1941.

[3] J. R. Pierce, "Limiting Stable Current in Electron Beams in the Presence of Ions," *Journal of Applied Physics*, v. 15, n. 10, pp. 721–726; 1944.

6. *Appendix*

6.1 SOLUTION OF EQUATION INVOLVED IN ION FLOW

Equation (4), in Section 1, may be presented as follows:

$$f(X_0)=c\int_{X_0}^{d}\frac{dX}{[f(X)-f(X_0)]^{\frac{1}{2}}}. \tag{16}$$

Where $f(X)$ is monotonically increasing to avoid an imaginary solution, we may take $y=f(X)$ without error in the following:
Let

$$X=g(y).$$

Then

$$dX=g'(y)dy.$$

Substitute this in the above integral, and let

$$y_0=f(X_0),\ L=f(d),\ \text{or}\ d=g(L).$$

Then

$$y_0=c\int_{y_0}^{L}\frac{g'(y)dy}{(y-y_0)^{\frac{1}{2}}}, \tag{17}$$

which is to be solved for $g'(y)$.

For the solution of this, take the formula[4,5]

$$\pi=\int^{y}\frac{dy_0}{[(y-y_0)(y_0-\eta)]^{\frac{1}{2}}}.$$

Multiply through by $g'(y)$:

$$\pi g'(y)=\int_{\eta}^{y}\frac{g'(y)\,dy_0}{[(y-y_0)(y_0-\eta)]^{\frac{1}{2}}}.$$

As $g'(y)$ is constant with respect to the variable y_0, it can go under integral sign. Integrate both sides with respect to y from η to L;

$$\pi[g(L)-g(\eta)]=\int_{\eta}^{L}dy\int_{\eta}^{y}\frac{g'(y)\,dy_0}{[(y-y_0)(y_0-\eta)]^{\frac{1}{2}}}.$$

Change the order of integration;

$$\int_{\eta}^{L}dy_0\int_{y_0}^{L}\frac{g'(y)\,dy}{[(y-y_0)(y_0-\eta)]^{\frac{1}{2}}}.$$

For the first integration take out the term independent of y;

$$\pi[g(L)-g(\eta)]=\int_{\eta}^{L}\frac{dy_0}{(y_0-\eta)^{\frac{1}{2}}}\int_{y_0}^{L}\frac{g(y)\,dy}{(y-y_0)^{\frac{1}{2}}}.$$

But the second term of the product is precisely y_0/c, or the unknown function g has been taken out of the integral into an explicit form on the left.

Then,[6]

$$\pi[g(L)-g(\eta)]=\frac{1}{c}\int_{\eta}^{L}\frac{y_0\,dy_0}{(y_0-\eta)^{\frac{1}{2}}}$$

$$=\frac{1}{c}\left[\frac{2(y_0+2\eta)}{3}(y_0-\eta)^{\frac{1}{2}}\right]_{\eta}^{L}$$

or

$$c\pi[g(L)-g(\eta)]=\frac{2(L+2\eta)(L-\eta)^{\frac{1}{2}}}{3}.$$

As η is a general variable, it may be replaced by y:

$$3c\pi[g(L)-g(y)]=2(L+2y)(L-y)^{\frac{1}{2}}.$$

Since $X=g(y)$, and $d=g(L)$,

$$3c\pi(d-X)=2(L+2y)(L-y)^{\frac{1}{2}}.$$

The solution of this equation is $y=f(X)$, where $y=L$ when $X=d$. Only this of the three possible solutions satisfies the original integral equation. We had originally in (4), $\rho=f(X)$, or we may write the solution as

$$3\pi c(d-X)=2(L+2\rho)(L-\rho)^{\frac{1}{2}}. \tag{6}$$

6.2 MATHEMATICAL SOLUTION IN PARAMETRIC FORM

Given the solution

$$C(d-X)=(L+2y)(L-y)^{\frac{1}{2}}, \tag{18}$$

where $C=3\pi c/2$, and c is the original constant specified in the derivation in (5); and given that C, d, and $\left(\frac{dy}{dX}\right)_{X=0}=m$ are known, where m is the gradient at the aperture in ions per centimeter, the problem is to obtain the value of L which satisfies the equation, where *L is the value of y at $X=d$.*

Let

$$\bar{x}=C(d-X),\ \text{or}\ \bar{x}=Cd, \tag{19}$$

if we work consistently at $X=0$ as we do in the remainder of this work. Then,

$$\frac{d\bar{x}}{dy}=-\frac{C}{m}\ \text{at}\ X=0. \tag{20}$$

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[4] Whittaker and Watson, "Modern Analysis," 4th Edition, Cambridge University Press, London; p. 229. Substitute $\mu=\frac{1}{2}$.

[5] Burington, "Mathematical Tables," 2nd Edition, Handbook Publishers, Inc., Sandusky, Ohio; section 158, p. 70. Take $+\pi$, as the integral is $+$ everywhere.

[6] Reference 5; Section 61, p. 61.

Let $M=-\frac{C}{m}$;

$$\begin{aligned}\bar{x}^2&=(L+2y)^2\,(L-y),\\ &=L^3(1+2v)^2\,(1-v),\text{ where } v=\frac{y}{L},\\ &=L\,(1+3v-4v^3).\end{aligned}\qquad(21)$$

Let $v=\cos\varphi$;

$$\bar{x}^2=L^3[1-\cos(3\varphi)]=2L^3\sin^2\left(\frac{3\varphi}{2}\right),\qquad(22)$$

since $-\cos(3\varphi)=(3\cos\varphi-4\cos^3\varphi)$.
Now

$$L^3=\frac{\bar{x}^2}{2\sin^2\left(\frac{3\varphi}{2}\right)},$$

or

$$L^3=\frac{C^2d^2}{2\sin^2\left(\frac{3\varphi}{2}\right)}\text{ at } X=0.$$

Then

$$L=(\tfrac{1}{2})^{\frac{1}{3}}\left[\frac{Cd}{\sin\left(\frac{3\varphi}{2}\right)}\right]^{\frac{2}{3}},\qquad(23)$$

which is the basic equation used in the numerical example in Section 2. Differentiate the first form in (22) with respect to y;

$$2\bar{x}\frac{d\bar{x}}{dy}=6L^3\left[\sin\left(\frac{3\varphi}{2}\right)\right]\left[\cos\left(\frac{3\varphi}{2}\right)\right]\frac{d\varphi}{dy}.\qquad(24)$$

But $y=L\cos\varphi$, so $1=-L\sin\varphi\frac{d\varphi}{dy}$, or,

$$2\bar{x}M=-6L^2\frac{\left[\sin\left(\frac{3\varphi}{2}\right)\right]\left[\cos\left(\frac{3\varphi}{2}\right)\right]}{\sin\varphi}.$$

Then,

$$\bar{x}^3M^3=-27L^6\frac{\left[\sin^3\left(\frac{3\varphi}{2}\right)\right]\left[\cos^3\left(\frac{3\varphi}{2}\right)\right]}{\sin^3\varphi}.\qquad(25)$$

But from (22),

$$\bar{x}^4=4L^6\sin^4\left(\frac{3\varphi}{2}\right).\qquad(26)$$

Eliminating L^6 between (25) and (26),

$$-\frac{4M^3}{27\bar{x}}=\frac{\cos^3\left(\frac{3\varphi}{2}\right)}{\sin^3\varphi\sin\left(\frac{3\varphi}{2}\right)}.$$

If we let

$$F(\varphi)=\frac{\cos^3\left(\frac{3\varphi}{2}\right)}{\sin^3\varphi\sin\left(\frac{3\varphi}{2}\right)},\qquad(27)$$

we can plot $F(\varphi)$ against φ. This has been done in Figs. 5 and 6.
Also,

$$F(\varphi)=-\frac{4M^3}{27\bar{x}},$$

or

$$F(\varphi)=\frac{4C^2}{27dm^3}.\qquad(28)$$

This is the basic equation from which $F(\varphi)$ is found. The value of the constant C is determined by voltage, current, and pressure in a manner shown in Section 6.3. When $F(\varphi)$ is known, then φ may be found in accordance with a plot of (27). With φ known, the value of L, the ion density at the back plate, may be found from (23).

6.3 Derivation of Constants

The constants of (19), (23), and (28) of Section 6.2 may be found as follows:

$$C=\frac{3\pi c}{2}=\frac{3\pi}{2}\cdot\frac{n}{K},$$

where n is the number of ions produced per second and K is the factor to multiply by the square root of the difference of the number of ions to give ion velocity in centimeters per second.

Now[7]

$$n=\frac{200pi}{Ve},$$

where p is in millimeters of mercury, i is in electrostatic units of current, V is in electrostatic units of voltage, and e is in electrostatic units of charge for an electron. Or

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[7] W. H. Bennett, "Magnetically Self Focusing Streams," *Physical Review*, v. 45, n. 12, pp. 890–897; 1934.

$$n = 3.75 \cdot 10^{23} \cdot \frac{pi}{V},$$

where p is in millimeters of mercury, i is in amperes, and V is in volts.

To find K, consider

$$v = 5.97 \cdot 10^7 V^{\frac{1}{2}} \left(\frac{m_e}{m_i}\right)^{\frac{1}{2}} \text{ centimeter per second,}$$

where V is in volts, and m_e/m_i is the ratio of the mass of the electron to the mass of ions involved (assuming essentially oxygen) $= \frac{1}{16} \times 1824$;

$$v = 3.49 \cdot 10^5 V^{\frac{1}{2}} \text{ centimeter per second.}$$

Or,

$$v = 6.04 \cdot 10^6 (\rho' +)^{\frac{1}{2}} \text{ centimeter per second,}$$

where $\rho' +$ is in electrostatic units of $+$ charge per centimeter (factor of $\sqrt{300}$ to convert V above from volts to statvolts). Then,

$$v = 1.32 \cdot 10^2 (\rho +)^{\frac{1}{2}} \text{ centimeter per second,}$$

where $\rho +$ is the number of $+$ ions per centimeter (factor of $(4.80 \cdot 10^{-10})^{\frac{1}{2}}$, singly ionized).

Or,

$$K = 1.32 \cdot 10^2.$$

Consequently,

$$C = 1.34 \cdot 10^{22} \cdot \frac{pi}{V}.$$

This is the C of (18), and appears in (8).

The constant of (28) and therefore of (9) is

$$\frac{4C^2}{27\left(\dfrac{1}{4.80 \cdot 10^{-10} \cdot 300}\right)^3},$$

where the gradient m is expressed in volts per centimeter, rather than ions per centimeter.

This is

$$7.96 \cdot 10^{22} \cdot \frac{p^2 i^2}{V^2}.$$

The constant of (23) and of (10) is

$$\left(\frac{C^2}{2}\right)^{\frac{1}{3}},$$

which is

$$4.47 \cdot 10^{14} \left(\frac{pi}{V}\right)^{\frac{2}{3}}.$$

Telephone Statistics of the World*

Countries	Date of Statistics	Telephones		Telephone Wire		Telegraph Wire		Conversations		Telegrams		Telephones in Large Cities		
		Thousands	Per 100 Pop.	Thousands of Miles	Per 100 Pop.	Thousands of Miles	Per 100 Pop.	Millions	Per Capita Avg.	Thousands	Per Capita Avg.	Exchange Area	Telephones Thousands	Per 100 Pop.
NORTH AMERICA:														
United States	Jan. 1, 1946	27867.0	21.0	110700	83.5	2260	1.7	36765	284.5	235000	1.8	New York City†	2002.3	27.8
Canada	Jan. 1, 1946	1692.2	14.4	6058	51.6	379	3.2	2980	254.7	15650	1.3	Montreal (1-1-46)	242.8	20.5
Mexico	Jan. 1, 1946	217.0	1.0	885	4.1	—	—	575	26.8	—	—	Mexico City	117.3	6.8
Cuba	Jan. 1, 1946	81.1	1.6	341	6.8	15	0.3	463	93.2	—	—	Havana	58.0	8.1
Puerto Rico	Jan. 1, 1946	25.8	1.3	76	3.7	—	—	58	28.5	—	—	—	—	—
Total	Jan. 1, 1946	30100.0	15.4	—	—	—	—	—	—	—	—	—	—	—
SOUTH AMERICA:														
Argentina	Jan. 1, 1946	571.0	4.0	3050	21.3	165	1.2	—	—	—	—	Buenos Aires (1-1-45)	346.7	9.7
Bolivia	Jan. 1, 1945	7.6	0.2	15	0.4	9	0.3	—	—	—	—	—	—	—
Brazil	Jan. 1, 1943	331.0	0.8	1360	3.2	114	0.3	1700	40.3	—	—	Rio de Janeiro	131.4	6.6
Chile	Jan. 1, 1946	109.5	2.0	382	7.1	—	—	369	68.8	—	—	Santiago	59.4	5.9
Colombia	Jan. 1, 1944	47.1	0.5	200	2.0	—	—	—	—	—	—	Bogota	16.5	4.1
Ecuador	Jan. 1, 1945	8.6	0.3	10	0.3	7	0.2	—	—	—	—	Quito	3.8	2.2
Paraguay	Jan. 1, 1945	4.2	0.4	11	1.0	4	0.3	—	—	—	—	—	—	—
Peru	Jan. 1, 1946	39.6	0.5	138	1.8	18	0.2	—	—	—	—	Lima (1-1-45)	26.1	4.7
Uruguay	Jan. 1, 1942	57.8	2.7	187	8.6	7	0.3	160	73.5	—	—	Montevideo	42.1	5.9
Venezuela	Jan. 1, 1942	36.1	1.0	120	3.3	8	0.2	214	58.6	—	—	Caracas	25.8	8.1
Guianas	Jan. 1, 1946	3.6	0.6	10	1.7	—	—	—	—	—	—	—	—	—
Total	Jan. 1, 1946	1290.0	1.3	—	—	—	—	—	—	—	—	—	—	—
EUROPE:														
Belgium	Jan. 1, 1946	379.6	4.5	—	—	—	—	289	34.3	—	—	Brussels	138.7	13.7
Bulgaria	Jan. 1, 1946	44.9	0.6	—	—	—	—	—	—	—	—	—	—	—
Denmark	Jan. 1, 1946	567.3	14.2	1630	41.0	—	—	960	241.7	—	—	Copenhagen	265.5	26.3
Eire	Jan. 1, 1946	55.1	1.9	194	6.6	20	0.7	56	19.0	2607	0.9	Dublin	31.7	6.5
Finland	Jan. 1, 1946	243.7	6.2	—	—	—	—	—	—	—	—	—	—	—
France	Jan. 1, 1946	1879.5	4.7	—	—	—	—	1358	34.2	40312	1.0	—	—	—
Great Britain	Mar. 31, 1945	3925.0	8.2	18500	38.5	—	—	—	—	—	—	—	—	—
Hungary	Jan. 1, 1944	256.9	1.7	580	3.9	—	—	340	22.8	—	—	—	—	—
Norway	Jun. 30, 1944	327.0	10.9	974	32.4	13	0.4	433	144.5	7717	2.6	Oslo (6-30-45)	95.6	22.0
Portugal	Jan. 1, 1946	97.7	1.2	210	2.6	—	—	—	—	—	—	Lisbon	42.0	5.7
Russia	Jan. 1, 1939	1272.5	0.8	2000	1.2	—	—	—	—	—	—	Moscow (1-1-36)	144.7	3.5
Spain	Jan. 1, 1946	447.2	1.6	1141	4.2	—	—	997	36.8	—	—	Madrid	95.4	7.9
Sweden	Jan. 1, 1945	1168.1	17.7	3929	59.6	8	0.1	1596	243.3	6000	0.9	Stockholm	310.5	38.4
Switzerland	Jan. 1, 1946	645.4	14.7	1950	44.3	27	0.6	543	124.0	2295	0.5	Zurich	102.6	27.4
Total	Jan. 1, 1946	16980.0	2.9	—	—	—	—	—	—	—	—	—	—	—
ASIA:														
British India	Jan. 1, 1946	118.6	0.03	776	0.2	413	0.1	—	—	—	—	Calcutta (3-31-45)	28.5	1.3
Total	Jan. 1, 1946	1500.0	0.1	—	—	—	—	—	—	—	—	—	—	—
AFRICA:														
Union of South Africa	Mar. 31, 1945	275.0	2.5	1165	10.4	19	0.2	380	34.1	9513	0.9	Johannesburg	72.1	11.0
Total	Jan. 1, 1946	430.0	0.2	—	—	—	—	—	—	—	—	—	—	—
OCEANIA:														
Australia	Jun. 30, 1944	799.7	10.9	3670	50.2	200	2.7	715	98.4	34721	4.8	—	—	—
Hawaii	Jan. 1, 1946	69.0	13.8	201	40.3	—	—	—	—	—	—	Honolulu	39.1	18.2
New Zealand	Mar. 31, 1946	265.8	15.6	—	—	—	—	—	—	—	—	—	—	—
Total	Jan. 1, 1946	1200.0	1.1	—	—	—	—	—	—	—	—	—	—	—
TOTAL WORLD	Jan. 1, 1946	51500.0	2.2	—	—	—	—	—	—	—	—	—	—	—

* Compiled by Chief Statistician's Division, American Telephone and Telegraph Company, and issued under date of September 10, 1946. Owing to conditions resulting from the war, official data for recent dates are not available for many countries. The above statistics were the latest available from authentic sources, on August 31, 1946. Totals for the world and geographical areas are partly estimated.

Grateful acknowledgment is hereby made of the courteous cooperation in the preparation of these statistics given by officials of private and governmental telephone and telegraph organizations and by representatives of the United States Departments of State and Commerce.

† The telephone development (telephones expressed in thousands) of other representative cities in the United States was, on January 1, 1946:

Chicago, Ill.	1204.5 telephones, or 34.5 per 100 population	Milwaukee, Wis.	215.6 telephones, or 27.2 per 100 population
Los Angeles, Cal.	591.0 " " 32.4 " " "	Minneapolis, Minn.	198.2 " " 35.8 " " "
Cleveland, Ohio	375.0 " " 31.2 " " "	Seattle, Wash.	190.3 " " 33.2 " " "
Washington, D. C.	364.4 " " 39.9 " " "	Denver, Col.	134.7 " " 36.4 " " "
San Francisco, Cal.	344.0 " " 43.3 " " "	Hartford, Conn.	90.8 " " 33.8 " " "
Boston, Mass.	232.0 " " 33.2 " " "	Omaha, Neb.	84.3 " " 31.2 " "

C.C.I.F. Honors Two Pioneers

AT THE 14th Plenary Assembly of the Comité Consultatif International Téléphonique (C.C.I.F.) at Montreux, Switzerland, on October 26, 1946, Mr. S. Rynning-Tønnessen, Director-General of the Norwegian Telegraph Administration, proposed the election as Honorary Members of Dr. A. Muri, former Director-General of the Swiss Posts, Telegraphs, and Telephones, and Sir Frank Gill, Chairman of the Board of Standard Telephones and Cables Ltd.

Mr. Rynning-Tønnessen said that some of the founders of the C.C.I.F. were still active, and in behalf of their colleagues and friends he wished to express respect, appreciation, and thanks to two of these pioneers, Sir Frank Gill and Dr. Muri.

Dr. Muri had attended most of the meetings, and has been greatly honored for his eminent skill and for his broadminded and tactful handling of C.C.I.F. questions and personalities. Sir Frank Gill has been honored in the many countries where he has interested himself in the peoples and their problems. He has been in a leading position in telephony for about 50 years, and has been concerned with all the problems of international telephony. It was largely due to his efforts that the Comité Consultatif International Téléphonique was formed, and he is still as enthusiastic as he was 25 years ago. The younger generation, consequently, wished to show these two gentlemen, in some way, how much they appreciated them. He therefore proposed that the 14th Plenary Assembly elect them as Honorary Members.

The proposal having been received with acclamation, Dr. F. Hess, Director-General of the Swiss Posts, Telegraphs, and Telephones, and President of the 14th Plenary Assembly, said that the warm applause of the delegates showed the unanimity of their desire to elect Sir Frank Gill and Dr. Muri as Honorary Members of the Comité Consultatif International Téléphonique. He cordially congratulated them both in the name of the Assembly.

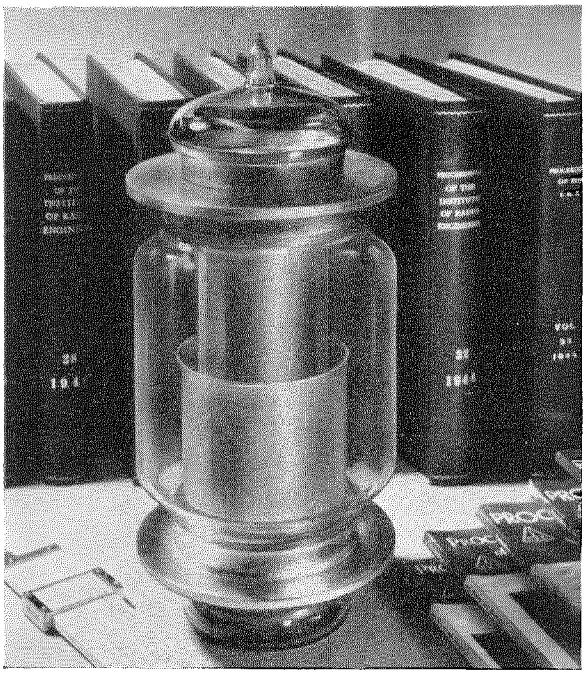
Recent Telecommunication Developments

BELL TELEPHONE MANUFACTURING COMPANY MAKES FIRST SERVICE AWARDS SINCE 1938—The annual service awards dinners, at which Bell Telephone Manufacturing Company, Antwerp, recognizes those employees who have served the company for 25 years or longer, were interrupted after 1938 by the war. On September 7, 1946, the tradition was resumed to honor 437 employees at a dinner in the famous Antwerp Bourse, or stock exchange, in which only outstanding State functions are held.

His Excellency C. Huysmans, Prime Minister of Belgium; His Excellency E. Rongvaux, Minister of Communications; Mr. Malderez, Secretary General of the Communications Ministry; Mr. K. Peeters, Head of the Cabinet of the Ministry of Labor and Social Security; Mr. J. Caers, Director General of the Labor Ministry, and Mr. W. Eekeleers, Deputy Burgomaster of Antwerp, represented the Government.

The Belgium Government decorates persons who have served industry continuously for 25 years or more, and awards were conferred on those who have completed such service since 1938: 25 Golden Palms and 35 Golden Medals of the Order of the Crown for 45 and 40 years, respectively, of service; 7 Golden Medals of the Order of Leopold II, 35 years; and Industrial Decorations numbering 103 to Workmen for 35 years, 205 for 25 years; and 62 to Clerks for 30 years.

Managing Director Leo Van Dyck, wearer of the Order of the Crown, and his staff of department heads awarded breloques to 349 employees who completed their 25th years during the war. These Bell Telephone Company awards, worn as watch charms by men and on necklaces by women, are a badge of industrial distinction.



R.M.S. *Queen Elizabeth* Sets Traffic Record—In October, 1946, on her maiden voyage as a luxury liner from Southampton to New York and return, *R.M.S. Queen Elizabeth* handled a total of 131 600 words of passenger and press radiotelegraph traffic to stations on both sides of the Atlantic, often to East and West simultaneously. Both manual and automatic transmission using speeds up to 100 words per minute were employed.

An all-time record for the highest volume of private radiotelegraph traffic exchanged by any one ship with Great Britain was established on the outward voyage, lasting less than five days, when 66 306 words were passed through British Post Office shore wireless stations. This exceeded the previous record of 48 004 words held by *R.M.S. Queen Mary.*

At the same time, 607 radiotelephone calls were put through to and from telephone subscribers in various countries, including South Africa and Australia. Also, 32 separate programs were transmitted to broadcasting systems throughout the world.

In addition to the above figures there was, of course, the large number of words transmitted and received in normal navigational messages and as "copy" for the daily newspapers published on board.

International Marine Radio Company (London) installed and operates the radio equipment on the *Queen Elizabeth* which for six years before her introduction to peacetime travelers was constantly engaged in transporting war personnel and supplies.

* * *

Vacuum Capacitors—Fixed capacitors for use in high-frequency transmitters of high power are now available from Federal Telephone and Radio Corporation. The electrodes consist of two coaxial copper cylinders mounted on ring seals similar to those used in some high-power vacuum tubes. A glass envelope and ends complete the enclosure which is evacuated.

Capacitors are made in 50- and 100-micromicrofarad sizes within tolerances of ±5 percent. For special purposes, larger capacitances and closer tolerances can be provided. Both are rated at a maximum peak value of 55 kilovolts and a root-mean-square current of 120 amperes.

With increasing frequency, the maximum permissible current usually limits the voltage that may be applied to a capacitor, resulting in the following ratings for these designs.

Peak Rating in Kilovolts	Frequency in Megacycles	
	50 $\mu\mu f$	100 $\mu\mu f$
20	50	33
40	25	16
55	18	12

Physically, the units are interchangeable, each being 15 inches tall and 6 inches in maximum diameter.

Contributors to This Issue

V. Belevitch

Vitold Belevitch was born at Helsingfors, Finland, on March 2, 1921. He studied until July, 1936, at the Notre-Dame de la Paix College at Namur. He then followed an engineering course at the University of Louvain and graduated in 1942 with an engineering degree.

Employed by the Bell Telephone Manufacturing Company at Antwerp in October, 1942, in the transmission department, he undertook various studies relating in particular to special transmission problems. In October, 1945, he received a doctor's degree in applied science from the University of Louvain.

* * *

E. M. Deloraine was born in Paris, France, on May 16, 1898. He received the B.S. degree, the Certificat de Mathématiques in 1918, and of Ingénieur diplomé de L'école de Physique et Chimie, a branch of Paris University, in 1920.

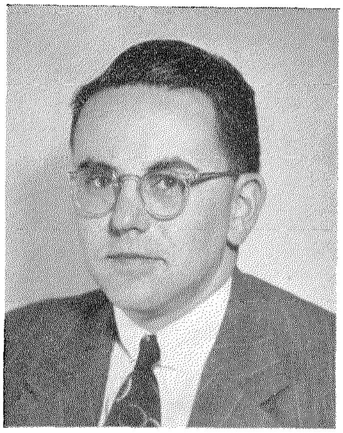
E. M. Deloraine

In 1917 he joined the French Army Signal Corps, and later engaged in research work at the Eiffel Tower. He became associated with the London engineering staff of the International Western Electric Company in 1921 and began technical work in connection with broadcasting at the experimental station 2WP. Until 1925 he was responsible for part of the developments in Great Britain in connection with the first transatlantic telephone circuit. He was made European technical director of International Standard Electric Corporation in 1933. From 1931 to 1937, Mr. Deloraine made important contributions in the development of ultra-high frequencies. He was also active in the advancement of high-power broadcasting, and established the Prague station with 120 kilowatts carrier, followed by the Budapest station. Mr. Deloraine was also successful in directing experiments in connection with automatic radio compasses for aircraft.

Mr. Deloraine came to the United States in 1941 to take charge of the organization of the laboratories unit for the Federal Telephone and Radio Corporation. In 1945, he was appointed president of International Telecommunication Laboratories, Inc. and in 1946 technical director of the International Telephone and Telegraph Corporation.

Mr. Deloraine was made a Chevalier of the Legion of Honor in 1938 for exceptional services to the Posts and Telegraphs Department of France, and he was elected vice-president of the French Institute of Radio Engineers in 1939. He has been a member of the International Consultative Committee of Long Distance Telephony since 1927, and is also a member of the French Astronomical Society. Mr. Deloraine is a fellow of the Institute of Radio Engineers and was its vice-president for 1946.

* * *



James S. Engel

James S. Engel was born in Brooklyn, New York, on November 18, 1920. He received the B.Sc. degree in electrical engineering in 1942 from Massachusetts Institute of Technology. Since 1942, he has been engaged in engineering work on aerial navigation equipment for Federal Telephone and Radio Corporation.

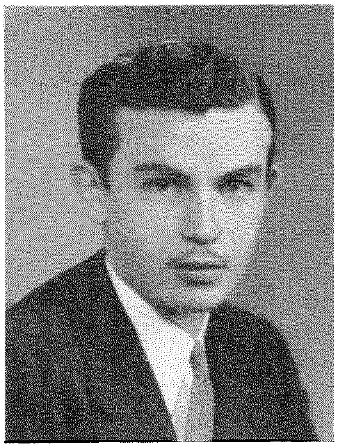
Mr. Engel holds associate memberships in the Institute of Radio Engineers and the American Institute of Electrical Engineers.

* * *

Leonard Everett was born on October 5, 1908 in Shanghai, China. He received from Stanford University an A.B. degree in engineering in 1931 and the E.E. degree in 1933.

From 1934 to 1940, he was employed as a sales engineer by the China Elec-

Leonard Everett



Lester M. Field

tric Company, Ltd. in Hong Kong and Shanghai. The following year was spent as a radio engineer for the Intercontinent Corporation in Rangoon, Burma.

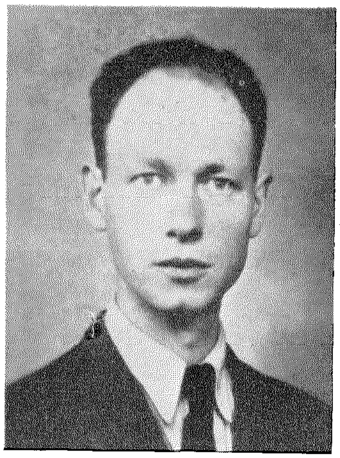
In March, 1942, he returned to the U.S.A. and joined the engineering staff of Federal Telephone and Radio Corporation.

Mr. Everett is a senior member of the Institute of Radio Engineers.

• • •

Lester M. Field was born on February 9, 1918, at Chicago, Illinois. He received the B.S. degree from Purdue University in 1939, and the Ph.D. degree from Stanford University in 1944.

He was acting instructor in 1941, and acting assistant professor from 1942 to 1944 in electrical engineering at Stanford University. In 1944, he joined Bell Telephone Laboratories as a member of the magnetron development group and later the electron dynamics group of the physical research department. In September, 1946, Dr. Field returned to Stanford University as acting associate professor of electrical engineering.



A. Fromageot

He is a member of the American Physical Society, Tau Beta Pi, and an associate member of Sigma Xi.

• • •

Antoine Fromageot was born on October 27, 1910, in Paris, France. He studied at the Ecole Polytechnique until 1932. In 1934, he received the degree of electrical engineer at the Ecole Supérieure d'Electricité (Paris), and spent the next year at the Cavendish Laboratory of Cambridge University.

Mr. Fromageot joined the Société Anonyme Lignes Télégraphiques et Téléphoniques in 1936; he was later placed in charge of development work related to coaxial and 12-channel carrier cables. In 1943, he was transferred to the Laboratoire Central de Télécommunications, where he is now directing development work on carrier telephony.

• • •

Robert Helm was born in Topeka, Kansas, on March 23, 1919. Stanford University conferred on him the A.B. degree in electrical engineering in 1940, and the E.E. degree in 1942.

After serving for a part of 1942 on the staff of Litton Engineering Laboratories, he became an engineer in the vacuum-tube division of Federal Telephone and Radio Corporation. In 1944, he rejoined Litton Engineering Laboratories as a research engineer on vacuum tubes.

Mr. Helm is a member of Sigma Xi, Phi Beta Kappa, Tau Beta Phi, and Kappa Sigma.

• • •

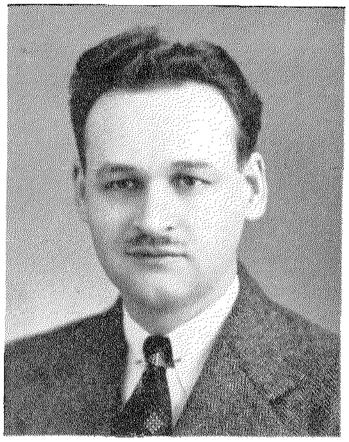
F. J. Mann was born in Hoboken, New Jersey, on April 27, 1906. He received the B.S. degree from Middlebury College in 1933. Postgraduate studies were in communications and journalism, the latter at Columbia University School of Journalism.

F. J. Mann

From 1925 to 1927, he was on the staff of *Radio Broadcast* magazine. In 1928, he was appointed radio editor of *Science and Invention.* He served as a marine radio operator before completing college and after his post-graduate work he became a radio operator at broadcast station WCAX in Burlington, Vermont. From 1934 to 1940, he was engaged in newspaper work and free-lance writing. Mr. Mann then became assistant to the director of technical publications of Bendix Radio.

Joining the *Electrical Communication* staff as a writer in 1942, he is now managing editor of the publication and assistant editor of the handbook, *Reference Data for Radio Engineers.*

A. A. New



Benjamin Parzen

Mr. Mann is a member of the honorary journalism fraternity, Pi Delta Epsilon, and of the Institute of Radio Engineers.

* * *

A. A. **New** received from Leeds University the B.Sc. degree in chemistry in 1922, B.Sc. Hons. in colour chemistry in 1923, and the M.Sc. degree in 1924.

He joined Standard Telephones and Cables in 1924 to do research on dyeing of yarn used for electrical purposes and later studied insulating oils. He was senior assistant chemist in the chemical laboratory at Woolwich in 1930 when he was transferred to research on fibrous dielectrics and placed in charge of optical work in the combined chemical, physical, and metallurgical laboratory, later known as the materials laboratory, of which he was appointed head in 1937. In 1946, he was transferred to Standard Telecommunication Laboratories in the same capacity.

Mr. New is a fellow of the Royal Institute of Chemistry and of the Chemical Society, an associate of the Institute of Physics, and holds memberships in the Society of the Chemical Industry, Textile Institute, Society of Dyers and Colourists, Ceramic Society, and British Rheologists Club.

* * *

Benjamin Parzen was born on April 5, 1913 in Poland. He received the B.S. degree in engineering from the College of the City of New York in 1936.

During 1936 and 1937, he was a radio inspector for the Federal Communications Commission. From 1938 to 1944, he served as a civilian engineer for the United States Navy. He joined Federal Telecommunication Laboratories in 1944 and is now a senior engineer.

Mr. Parzen is a member of the American Institute of Electrical Engineers and is licensed as a professional engineer by New York State.

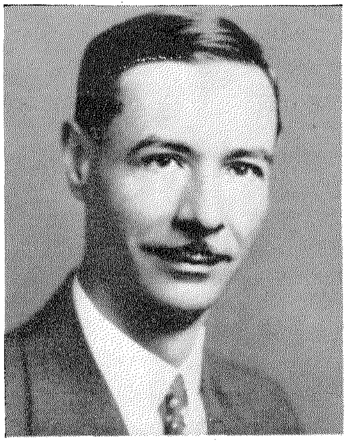
* * *

Haraden Pratt was born in San Francisco, California, on July 18, 1891. He received a B.S. degree from the University of California in 1914.

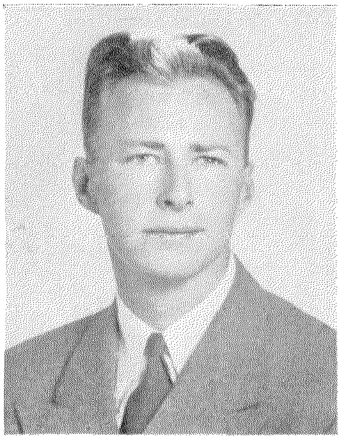
Starting his radio career as an amateur in 1906, Mr. Pratt became a commercial operator and installer for United Wireless Telegraph Company and Marconi Wireless Telegraph Company before completing his college training. After receiving his degree, he was engaged as an engineer by the Marconi Company.

From 1915 to 1920, he served as expert radio aide, U.S. Navy. From 1920 to 1923, he was an engineer for Federal Telegraph Company. He served Western Air Express on communication problems from 1925 to 1927 and during the following year was with the National Bureau of Standards.

In 1928, Mr. Pratt became chief engineer of Mackay Radio and Telegraph Company. He is now vice president, chief engineer, and a director of that company; vice president and chief engineer of American Cable and Radio Corporation, of All America Cables and Radio, Incorporated, and of Commercial Cable Company; vice president and a director of Federal Telephone and Radio Corporation and of Federal Telecommunication Laboratories; and a director of International Telecommunication Laboratories.



Haraden Pratt



Karl R. Spangenberg

The Institute of Radio Engineers presented its Medal of Honor to Mr. Pratt in 1944. He served as its president in 1938, as treasurer in 1941 and 1942, and as secretary from 1943 to date.

He is a fellow of the Institute of Radio Engineers and of the American Institute of Electrical Engineers.

Since 1927, he has attended a number of the major international conferences on radio.

* * *

Karl R. Spangenberg was born at Cleveland, Ohio, on April 9, 1910. He received the B.S. degree in electrical engineering in 1932, and the M.S. degree in electrical engineering in 1933, both from Case School of Applied Science; and the Ph.D. degree from Ohio State University in 1937.

Since 1937, Dr. Spangenberg has been a member of the faculty of the electrical engineering department of Stanford University, where he now holds the rank of professor of electrical

Arthur Van Dyck

engineering. During the war he was granted a leave of absence from Stanford University to serve as consultant to the Signal Corps and to work at the Radio Research Laboratory.

He is a member of the Institute of Radio Engineers, Tau Beta Pi, and Sigma Xi.

* * *

Arthur Van Dyck received from Yale University in 1911 a Ph.B. degree in electrical engineering.

He served on the engineering staffs of the National Electric Signalling Company, Westinghouse Electric and Manufacturing Company, and the Carnegie Institute of Technology. During World War I, he was expert radio aide for the United States Navy.

Since 1919, he has been with the Marconi Company and its successor, Radio Corporation of America, where he is assistant to the executive vice president in charge of the laboratories division. From 1943 to 1946, he served as Commander, U.S.N.R.

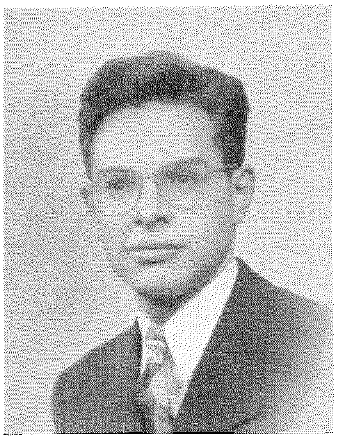
Mr. Van Dyck was awarded the Legion of Merit and the Naval Reserve Medal. He is a fellow and past president of the Institute of Radio Engineers, a fellow of the Radio Club of America, and a member of the Veteran Wireless Operators Association.

* * *

E. P. G. Wright joined the equipment engineering branch of the Western Electric Company in London in 1920, and in 1926 took charge of project engineering when the company became associated with the I.T.&T. System. In 1928, Mr. Wright was transferred to the I.T.&T. laboratories to work on switching problems. He was placed in charge of switching-system design for Standard Telephones and Cables in 1932 until the war when he was assigned to special duties on radio production and test-equipment design.

Mr. Wright served as a member of the British Telephone Technical Development Committee from its inception until 1939. He is a member of the Institution of Electrical Engineers and was awarded the Fahie Premium for papers on the bypath system and voice-frequency signalling and dialling.

E. P. G. Wright



A. Aaron Yalow

* * *

A. Aaron Yalow was born in Syracuse, New York, on September 18, 1919. He received an A.B. degree, magna cum laude, from Syracuse University in 1939. The University of Illinois granted him an M.S. degree in 1942 and a Ph.D. in physics in 1945.

He served as a graduate assistant in physics at the Univerity of Syracuse from 1939 to 1941, and at the University of Illinois during the following academic year. He then held a fellowship in physics until 1945.

Mr. Yalow joined the engineering staff of Federal Telecommunication Laboratories in 1945 and in 1946 became a physicist in the department of medical physics of Montefiore Hospital in New York City.

He is a member of the American Physical Society, Sigma Xi, Phi Beta Kappa, Sigma Pi Sigma, and Pi Mu Epsilon.

INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION

Associate Manufacturing and Sales Companies

UNITED STATES OF AMERICA

International Standard Electric Corporation, New York, New York

Federal Telephone and Radio Corporation, Newark, New Jersey

GREAT BRITAIN AND DOMINIONS

Standard Telephones and Cables, Limited, London, England

Branch Offices: Birmingham, Leeds, Manchester, England; Glasgow, Scotland; Dublin, Ireland; Cairo, Egypt; Calcutta, India; Johannesburg, South Africa.

Creed and Company, Limited, Croydon, England

International Marine Radio Company Limited, Liverpool, England

Kolster-Brandes Limited, Sidcup, England

Standard Telephones and Cables Pty. Limited, Sydney, Australia

Branch Offices: Melbourne, Australia; Wellington, New Zealand.

Silovac Electrical Products Pty. Limited, Sydney, Australia

New Zealand Electric Totalisators Limited, Wellington, New Zealand

Federal Electric Manufacturing Company, Ltd., Montreal, Canada

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Standard Electrica, S.A., Rio de Janeiro, Brazil

Compañía Standard Electric, S.A.C., Santiago, Chile

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Bell Telephone Manufacturing Company, Antwerp, Belgium

China Electric Company, Limited, Shanghai, China

Standard Electric Doms A Spolecnost, Prague, Czechoslovakia

Standard Electric Aktieselskab, Copenhagen, Denmark

Compagnie Générale de Constructions Téléphoniques, Paris, France

Le Matériel Téléphonique, Paris, France

Les Téléimprimeurs, Paris, France

Lignes Télégraphiques et Téléphoniques, Paris, France

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Standard Villamossági Részvény Társaság, Budapest, Hungary

Telefongyár R.T., Budapest, Hungary

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Societa Italiana Reti Telefoniche Interurbane, Milan, Italy

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Standard Telefon- og Kabelfabrik A/S, Oslo, Norway

Standard Electric Company w. Polsce Sp. z.O.O., Warsaw, Poland

Standard Electrica, Lisbon, Portugal

Standard Fabrica de Telefoane si Radio S.A., Bucharest, Rumania

Compañía Radio Aerea Maritima Española, Madrid, Spain

Standard Eléctrica, S.A., Madrid, Spain

Aktiebolaget Standard Radiofabrik, Stockholm, Sweden

Standard Telephone et Radio S.A., Zurich, Switzerland

Jugoslavensko Standard Electric Company Akcionarno Drustvo, Belgrade, Yugoslavia

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Telephone Operating Systems

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Compañía Telegráfico-Telefónica Comercial, Buenos Aires, Argentina

Compañía Telegráfico-Telefónica Del Plata, Buenos Aires, Argentina

Companhia Telefonica Paranaense S.A., Curitiba, Brazil

Companhia Telefonica Rio Grandense, Porto Alegre, Brazil

Compañía de Teléfonos de Chile, Santiago, Chile

Compañia Teléfonica de Magallanes S.A., Punta Arenas, Chile

Cuban Telephone Company, Havana, Cuba

Mexican Telephone and Telegraph Company, Mexico City, Mexico

Compañía Peruana de Teléfonos Limitada, Lima, Peru

Porto Rico Telephone Company, San Juan, Puerto Rico

Shanghai Telephone Company, Federal, Inc., U.S.A., Shanghai, China

Radiotelephone and Radiotelegraph Operating Companies

Compañía Internacional de Radio, Buenos Aires, Argentina

Compañía Internacional de Radio Boliviana, La Paz, Bolivia

Companhia Radio Internacional do Brasil, Rio de Janeiro, Brazil

Compañía Internacional de Radio, S.A., Santiago, Chile

Radio Corporation of Cuba, Havana, Cuba

Radio Corporation of Porto Rico, San Juan, Puerto Rico†

† Radiotelephone and Radio Broadcasting services.

Cable and Radio Telegraph Operating Companies

(Controlled by American Cable & Radio Corporation)

The Commercial Cable Company, New York, New York[1]

Mackay Radio and Telegraph Company, New York, New York[2]

All America Cables and Radio, Inc., New York, New York[3]

The Cuban All America Cables, Incorporated, Havana, Cuba[1]

Sociedad Anónima Radio Argentina, Buenos Aires, Argentina[4]

[1] Cable service. [2] International and Marine Radiotelegraph services. [3] Cable and Radiotelegraph services. [4] Radiotelegraph service.

Laboratories

International Telecommunication Laboratories, Inc., New York, New York

Federal Telecommunication Laboratories, Inc., New York, New York

Standard Telecommunication Laboratories Ltd., London, England

Laboratoire Central de Télécommunications, Paris, France

ELECTRICAL COMMUNICATION

June, 1947

Vol. 24—No. 2

Microwave Radio Relay Systems
Pulse-Time-Modulated Multiplex Radio Relay System
3000-Megacycle Frequency-Modulation Radio Link
Mobile Frequency-Modulation 30–44-Megacycle Equipment
Medium-Power Triode for 600 Megacycles
Triodes for 10-Kilowatt Frequency-Modulated Transmitters
Insertion Loss and Effective Phase Shift in Composite Filters
Portable Direct-Reading Traffic Recorder
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Errors in Spaced-Collector Direction-Finder Systems
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Survey of Radio Navigational Aids
Frequency-Shift Keying in the I.T.&T. System
Institute of Radio Engineers 1947 Convention

ELECTRICAL COMMUNICATION

Technical Journal of the
INTERNATIONAL TELEPHONE & TELEGRAPH CORPORATION
and Associate Companies

Volume 24 • Number 2—June 1947

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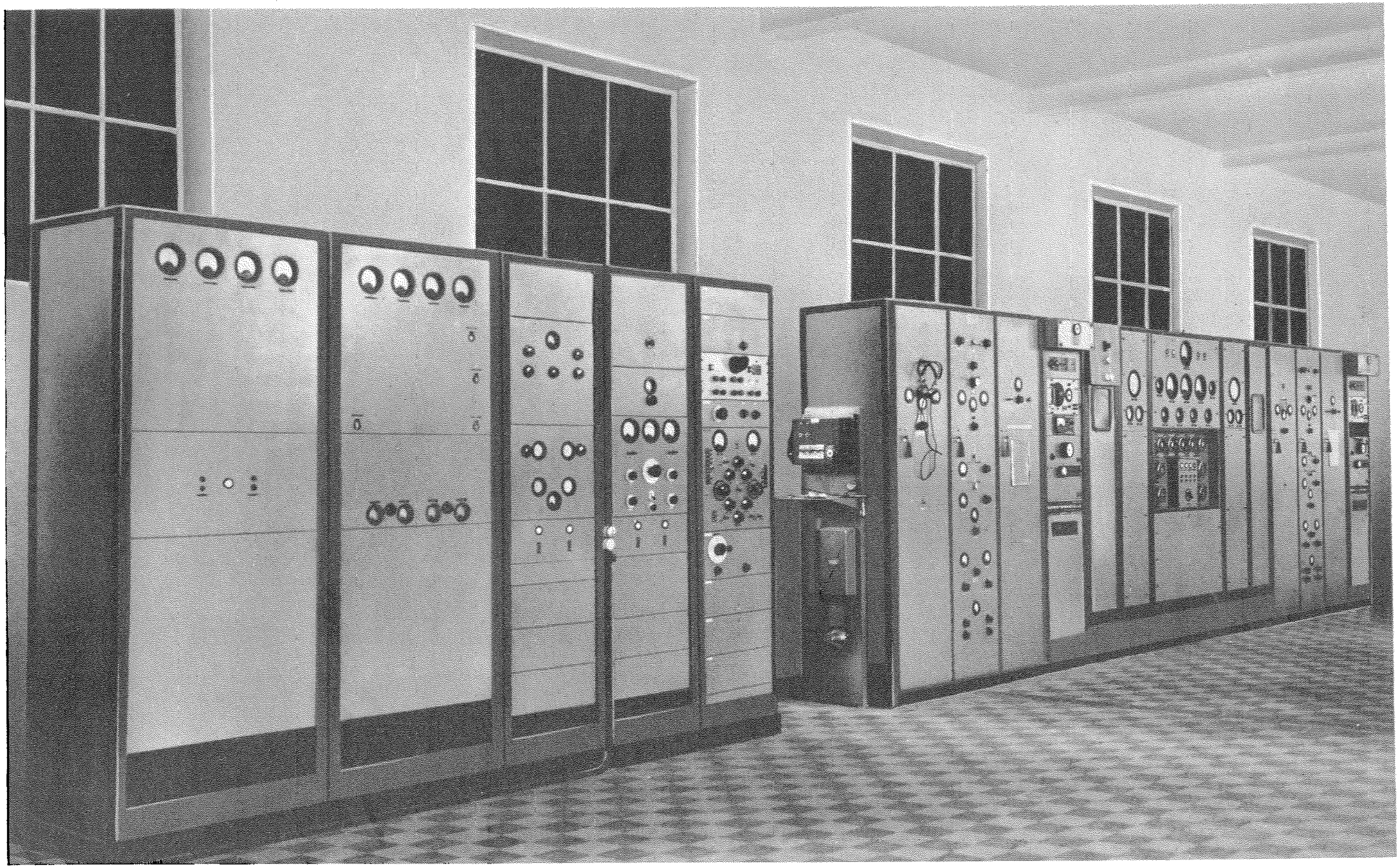

The single-sideband high-frequency transmitter in the illustration provides two telephone channels for the New York-Paris link and is installed at Pontoise, France. At the left is a 1-kilowatt driver and at the right is the 60-kilowatt output stage. The construction of this equipment by Laboratoire Central de Télécommunications is a logical result of its many years of work in this field.

Microwave Radio Relay Systems*

By E. LABIN

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VARIOUS technical problems involved in the design of microwave radio relay links are surveyed. Propagation characteristics, types of modulation, signal-to-noise conditions, and system engineering aspects are considered. The great value and the possible limitations of such links are pointed out.

* * *

The final goal of telecommunications often has been defined as being the ability of communicating with anyone, anywhere, any time. Ten years ago, this was still a dream, now it seems that it might come true. The scientific foundations for fulfilling our goal exist already. However, it will certainly be a long time before such a global communication network can be completed, and certainly social and economic difficulties will be at least as important as the technical factors.

The main progress that has made this dream possible is the advance accomplished recently in the extension of the radio-frequency range actually available to the engineer. The existing telecommunication network is essentially composed of wires and cables, extended by a relatively small number of fixed and mobile radio links. The fixed radio links were, until recently, almost all below 30 megacycles per second. In spite of the small density of this radio network, the ether below 30 megacycles is terribly crowded and there is practically no space available for additional services. The shortage of frequencies explains the constant efforts of all research laboratories to extend the range of practicably workable frequencies toward higher and higher values.

The laboratories of the I.T.&T. System recognized the importance of this question a long time ago, and in fact by 1932, an experimental microwave link was tested across the English Channel. But these experiments were isolated and far in advance of the general state of the art; so much so, that in 1938 the frequency range that we really knew how to handle did not extend above 500 megacycles. The accelerated developments during the war have changed this picture, and now the practical limit of usable frequencies has been extended up to at least 30,000 megacycles.

In the last ten years, it has not been possible to explore completely this vast range of frequencies. Much remains to be done and practically all telecommunication laboratories are busily developing new tools and various applications of microwaves. Aerial navigation and mobile links for communication with planes, cars, ships, and railroads are some of them. Microwaves are also invading a domain that wires and cables have previously filled. Many projects have been described for using microwave links with repeaters for long-range communication, television, telegraphy, and multichannel telephony. The purpose of this paper is to present a short general survey of the technical aspects of this last problem.

1. General

Microwave radio relay systems consist of a series of transmitters and receivers spaced 30 to 60 miles apart in direct visibility. On flat ground, towers are used to extend the horizon. The signal is retransmitted at each relay or repeater station, generally at a different frequency from the one at which it has been received. Relay systems of this sort have been proposed for television transmission or multichannel telephony. Essentially, it is hoped that relay radio systems not requiring physical connections between repeaters will be more economical than wire or cable connections and that the relatively greater facilities offered by radio for wide-frequency-band operation will open up new types of services. Most proposals now concern carrier frequencies between 1000 and 10,000 megacycles. There are two reasons for choosing such high frequencies.

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* Presented before meetings of Associazione Elettrotecnica Italiana in Rome, Italy, on November 5, 1946, and in Milan, Italy, on November 9, 1946. Reprinted from *Elettrotecnica*, June, 1947.

A. Higher frequencies have certain technical advantages.
B. Lower frequencies are indispensable for existing services.

It should, indeed, be clear that mobile services cannot, generally, make use of highly concentrated radiations. For equal directivity at the transmitter and receiver and for equal transmitted power, the received power is proportional to the square of the wavelength. In other words, for the same transmitted power, the power received for a comparable angular coverage would be one hundred times more at 100 megacycles than at 1000 megacycles. This is why mobile services rightly request priority for lower frequencies.

For fixed services, the situation is technically reversed, and because maximum directivity not only can be used but is even desirable, the useful power actually increases with frequency. Factors which must be considered are the propagation characteristics of waves between 1000 and 10,000 megacycles, various types of modulation, signal-to-noise ratio, and the over-all system aspects of microwave relay links.

2. *Propagation*

For fixed installations, the propagation characteristics can be expressed in a most useful manner by the attenuation between the transmitting point and the receiving point. If P_t designates the transmitted power and P_r the received power, the attenuation A is:

$$A=\frac{P_t}{P_r}=A_0\,A_1\,A_2\,A_3, \qquad (1)$$

where

A_0 = free-space factor,
A_1 = topographical factor,
A_2 = atmospheric factor, and
A_3 = absorption factor.

In (1), the attenuation has been expressed as a product of various factors. For convenience, these factors may be expressed in decibels, the total attenuation then being their sum.

2.1 Free-Space Factor

The free-space factor A_0 represents the attenuation that would exist in free space at a given distance between transmitter and receiver. It is a factor which can be calculated closely by the principles of geometrical optics. Essentially, it represents the loss resulting from the fact that the beam is not concentrated in an infinitely narrow angle. It is clear from Fig. 1, that A_0 is given by (2).

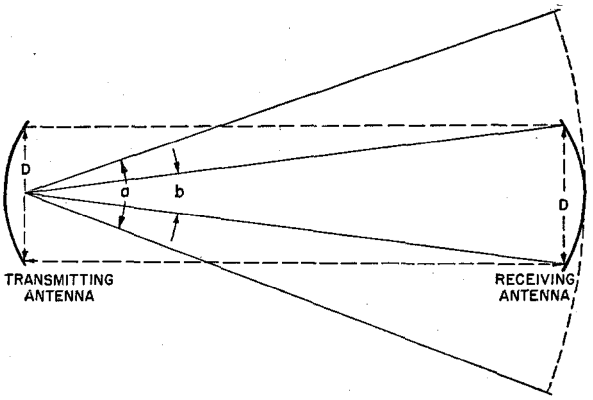

Fig. 1—Free-space attenuation is a function of the beam characteristics of the antennas. The angle in which the radiation is concentrated when using dipole radiators and parabolic reflectors is $a=1.7\ (\lambda/D)$. The angle in which the receiving antenna is viewed from the transmitting point is $b = D/L$, where L is the distance between transmitting and receiving antennas.

$$A_0=\left(\frac{a}{b}\right)^2. \qquad (2)$$

For sake of simplicity, axial symmetry is assumed. The angle a is the equivalent beam width, i.e., the width the beam would have if all the energy were radiated uniformly into that angle. It differs only slightly from the angle usually considered between the directions of half-power points. For a given type of antenna, the value of a can be obtained from diffraction theory of electromagnetic waves. If the antenna aperture is D, it can be shown that a is proportional to the ratio λ/D.

The exact numerical coefficient depends on the type of radiator, type of illumination, shape of reflectors, etc. For parabolic reflectors and dipole radiators, the angle a is given by

$$a=1.7\,\frac{\lambda}{D}\ \text{radians}, \qquad (3)$$

The angle b designates simply the angle under which the receiving antenna is viewed from the transmitting point and is essentially given by

$$b=\frac{D}{L}\text{ radians.} \tag{4}$$

Transposing (3) and (4) into (2) yields

$$A_0=3\,\frac{\lambda^2L^2}{D^4}. \tag{5}$$

This simple derivation of the free-space coefficient A_0 applies, of course, only when D is much larger than λ and when a is larger than b; and also assumes no fringing effects. It is interesting to note that $a=b$, which would mean no attenuation in free space, corresponding to conditions which can actually be met.

In most practical cases for distances of the order of 30 miles, A_0 will result in a power ratio of the order of 10^6 to 10^8 or of 60 to 80 decibels. The importance of using short wavelengths, i.e., small values of a, appears clearly from (2) or (5). It is also obvious from (5) that the most important single factor is the diameter D of the antennas used because it affects both angles a and b in the same proportion.

It is not possible to reduce the equivalent beam width a indefinitely. Two lower limits have to be considered: (1) The fluctuation of the position of the beam as a result of atmospheric conditions. This limit is probably below half a degree or 1/100 radian; and (2) the fluctuation of the position of the beam because of the limited mechanical rigidity of the antenna support. It seems likely that values of 0.03 to 0.06 radian would represent in most cases a good compromise between the costs of additional radio power and the additional costs of extremely rigid supporting structures.

2.2 TOPOGRAPHICAL FACTOR

Actually, the transmitter and receiver are not in free space. The presence of the earth changes considerably the simple propagation conditions of free space. When the transmitter is directly visible from the receiver over a smooth spherical earth, the coefficient A_1 can be calculated and many papers have been published dealing with this problem. A_1 in such conditions depends on the height of the transmitters and receivers above ground. A well-known interference pattern results from the reflection from the ground.

It is also well known that when the atmosphere is supposed to be "standard," the bending of the waves resulting from the variation of the index of refraction with altitude can be taken into account by using an effective radius for the earth's surface equal to 4/3 of the true radius. The theory of the factor A_1 is important especially at lower frequencies of the order of 30 to 300 megacycles. At these frequencies, it might be difficult to find a position of the transmitter and receiver that would make A_1 close to unity and, therefore, it is important to know what value it might take. For higher frequencies and narrow beam angles, it is not safe to assume that the earth is smooth; as soon as the link is not over flat ground, most of the formulas do not apply, and also the simple application of wave theories is often misleading.

As a compensation for these theoretical difficulties, the practical conditions are such for this frequency range that A_1 is generally close to unity if the height of the transmitting and receiving antennas are properly chosen and, therefore, the exact knowledge of A_1 is not too important. However, in special surroundings, such as a narrow valley or street, the factor A_1 could, of course, be quite different from unity.

Furthermore, from the point of view of the communication engineer, what is more important than the absolute value is the variation with time of the coefficient A_1. These changes are caused by variations of reflection coefficient of the ground. On land, variations are small because they result only from variations of conductivity at the reflection points caused by changes of humidity of the ground or presence of leaves or snow, etc. Such variations usually do not exceed 6 decibels. The variations might be larger if the reflecting region is over sea. It might move with the tides, but even so, the variations of A_1 would hardly be larger than 6 to 12 decibels, because the minima of the interference pattern resulting from reflection from the ground are seldom less than 12 decibels below the free-space conditions.

2.3 ATMOSPHERIC FACTOR

All that has been said before concerns standard atmospheric conditions. A standard atmosphere is defined as an atmosphere in which the refraction index varies steadily with altitude at a given rate. When the atmosphere is not standard, the propagation conditions might be greatly changed and the factor A_2 can reach very large values. Important theoretical and experimental investigations of variations of propagation with changes of climatic conditions in the lower atmosphere have been made recently. It has been found that variations of the coefficient A_2 by as much as 30 to 40 decibels are possible depending on a great variety of factors. Usually in these studies, the refractive state of the atmosphere is most conveniently described by means of a curve showing the relation between the so-called excess modified index of refraction M and the height h. M is given by

$$M = \left[n(h) + \frac{h}{r\text{-}1} \right] \cdot 10^6, \qquad (6)$$

where $n(h)$ is the actual index of refraction at height h above the ground, and r is the radius of the earth.

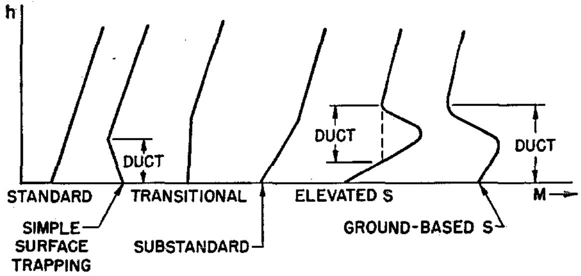

Fig. 2—Variations of refractive index of the atmosphere M with height h. The regions marked "duct" are capable of propagating waves over very long distances with little attenuation.

Fig. 2, taken from a report by I. Katz of the Radiation Laboratory of Massachusetts Institute of Technology, shows various possible aspects of the M curves. Various types of M curves correspond to various meteorological conditions. Only when the M curve is standard, or when it has a different slope but remains uniform, can the simple ray theory be applied to determine the value of the coefficient A_2. A slope variation of the M curve is equivalent to a bending of the whole interference pattern which determines the previous coefficient A_1.

When the M curve has a knee, the conditions become much more complicated and an actual duct is formed in the atmosphere, usually between the ground and the altitude at which the M curve changes its slope. These altitudes may vary from a few hundred to a few thousand feet. The waves can be trapped in the duct and be propagated for very long distances, but they might also never reach the desired receiving point if that point is not located in the right position with respect to the duct. Complete trapping is rather infrequent and, therefore, complete fading at the receiving point is seldom encountered but, as stated previously, variations up to 40 decibels have been observed at many frequencies. It is not obvious which frequencies are to be preferred from this point of view. It is true that, generally speaking, for a given duct position no trapping will occur at frequencies lower than a certain value but, the height of the duct being variable, this minimum frequency is also variable.

The relation between the types of M curves and meteorologic data are complex, but it seems that some very broad statements can be made. Irregularities in the M curve generally correspond to temperature inversions. But not all inversions will yield ducts. Often these conditions are realized when warm dry air blows out over a cool body of water. Similar effects can be obtained at night in summer over land. Generally speaking, when the air is strongly agitated or in windy weather, standard propagation conditions will prevail over land. In summer, and especially over sea, vagaries of propagation are more likely.

2.4 ABSORPTION FACTOR

The coefficient A_3 characterizes additional losses caused by true absorption in the atmosphere. For the factors previously considered, A_0, A_1, A_2, it is assumed that no absorption takes place in the atmosphere. This assumption is justified up to frequencies as high as 6000 megacycles. Fog, rain, or snow affect the propagation of frequencies at 3000 megacycles, but the losses encountered most of the time are negligible and do not represent more than 0.05 decibel per mile on the average. At wavelengths as short as 1 centimeter (30,000 megacycles), on the contrary, absorptions of 35 decibels per mile can be observed for heavy rainfalls. Most of the absorption effects result from the presence of water which

TABLE I

A Multiplexing	B Modulation of Each Channel	C Modulation of Radio-Frequency Carrier
a. Frequency Selection	a. Amplitude Modulation Single Sideband	a. Amplitude Modulation
b. Time Division	b. Frequency Modulation	b. Frequency or Phase Modulation
	c. Pulse-Amplitude Modulation	c. Amplitude Keying
	d. Pulse-Time Modulation	d. Frequency Keying
	e. Pulse-Code Modulation	e. Modulation *a* or *b* by a Subcarrier Itself Modulated by Methods *a* or *b*

has a strong absorption region at wavelengths around 1.25 centimeters (24,000 megacycles).

It is interesting to note that the factor A_3 represents a type of attenuation which varies exponentially with distance and which is, therefore, similar to what is encountered normally for propagation on wires or cables. As soon as factors of this type become important, they overshadow all others because they increase so rapidly with distance.

The practical conclusion concerning the various aspects of propagation conditions which have been outlined is that, in communication links, provisions for fading should be made taking into account variations of received power of the order of 30 to 40 decibels. It is possible that, in many cases, these variations could be considerably reduced by the use of two receivers at different heights. This method of space diversity has been tested only recently and it is difficult to estimate how successful it will be. Also, in spite of the abundance of information already available, it is still difficult to state how often fading conditions will be present for a given link and, therefore, what the probability of failure of the link might be. A 30-decibel margin in power is quite feasible and, therefore, the actual probability of failure could be relatively small, especially in cases where space diversity provides an additional protection. Some extreme cases of fading will probably be encountered where neither additional power nor space diversity can help.

3. *Modulation*

We are mainly concerned with multiplex telephony or telegraphy. The method of modulation best suited to the problem would be one which would require minimum transmitted power for a certain amount of noise present, minimum bandwidth, and would minimize difficulties due to the accumulation of distortions of each repeater. It is, of course, difficult to find a single figure of merit that would take into account all these various factors. Efforts have been made recently to arrive at a classification of various modulation methods by using the Hartley law relating the amount of intelligence transmitted, bandwidth used, and time required to transmit the intelligence, but even such efforts do not take into account the type of tubes available or the relative difficulty of avoiding distortions which might favor one type of modulation with respect to another. A list of the most promising methods will be given together with some information concerning the types that have been actually used.

Essentially three different operations have to be performed for the transmission of multiplex telephony:

A. Multiplexing of the various channels.

B. Transmission of the intelligence in each individual channel.

C. Modulation of the radio-frequency carrier with the complex signal formed by *A* and *B* above.

Each of the three operations described can be performed in various manners and a large number of systems can be obtained by combining the methods for each operation in various ways. Table I, which is far from complete and is limited to the most interesting types, will give an idea of the variety of systems possible.

In ordinary radio practice where the ether is crowded, one of the most important factors is economy of bandwidth. The same view certainly will apply to microwave operation but because of the unavoidable instability of the carriers used, the economical bandwidth is generally much larger than at lower frequencies. Even with the greatest of care, it is unlikely that the frequencies of the transmitter and receiver in a radio relay system could be adjusted and maintained stable within better than 0.01 percent. This means that at 5000 megacycles the bandwidth necessary to take care of instabilities alone would be at least 1 megacycle. In other words, the microwave equipments are inherently relatively wide band and, therefore, it seems reasonable to use modulation schemes which require bandwidths larger than strictly necessary for transmitting the intelligence but which make good use of these extended bandwidths by reducing distortion difficulties or by improving the signal-to-noise ratio.

Both frequency modulation and pulse-time modulation have similar advantages from these points of view. Although the choice between these two systems has been discussed at length, it seems obvious from Table I that the true problem is not what type of modulation of the radio frequency is to be used but rather to determine the most desirable multiplexing method. There are only two multiplexing systems: frequency selection where all channels are transmitted simultaneously and are seperated by frequency filters; or time division where the channels are transmitted successively.

Frequency selection is widely used for carrier transmission on wires and the same system should receive first consideration for radio. The main difficulty is that frequency selection requires extremely linear repeaters. To reduce cross talk below 60 decibels, the total over-all harmonic content between terminals has to be less than 0.1 percent. If 20 repeaters are to be put in tandem, each repeater should contribute less than 0.05 percent to the harmonic content over the signal frequency spectrum! It is not too difficult to meet such severe specification for wire communication with negative-feedback amplifiers. But until a similar tool is developed for radio repeaters, they will necessarily involve modulation and demodulation processes that introduce nonlinearities even if the radio carrier is frequency modulated.

One way out, which has been applied in the equipment developed for Western Union by Radio Corporation of America,[15] is to use a subcarrier. This system is a combination of the types *Aa*, *Ba*, *Ceb* of Table I.

For a limited number of repeaters, frequency or phase modulation is possible. It might be feasible to increase the number of repeaters by using a large amount of negative feedback. An experimental 12-channel equipment[19] of that type has been developed by Laboratoire Central de Télécommunications, for the French administration. It is now in operation between Paris and Montmorency.

Another possibility is to perform the multiplexing operation by time division and use pulses. For modulation of each channel, either pulse-amplitude or pulse-time modulation or the more complex telegraphic methods, such as pulse-code modulation, could be used. For modulation of the radio carrier, the choice remains between amplitude or frequency modulation. In any case, linearity requirements are much less severe. For time modulation, the linearity of the repeater is of no consequence at all for cross talk. A system using time-division multiplex for 24 channels has been developed by Federal Telecommunication Laboratories. It is a combination of the types *Ab*, *Bd*, *Cc*, of Table I. The main characteristics of the system are reproduced in the appendix. An experimental link with two repeaters is now in operation between New York and Trenton.

The total frequency band necessary might appear somewhat large but it should be noted that the system based on frequency selection with a subcarrier finally leads to quite similar frequency bands. Both systems take advantage of bandwidth to overcome the inherent difficulties of the design of linear repeaters.

Important improvements can be expected soon for both frequency-selection and time-division methods. It is difficult to say now which one is preferable and it seems most likely that both will find useful fields of application.

4. *Signal-to-Noise Ratio*

In all communication problems, the major question is finally: How large must the input power be to override all noise present and to achieve the output signal-to-noise ratio required by the type of service desired?

---

[15] Numbered references will be found on page 140.

At microwaves, barring interferences, the only appreciable sources of noise are thermal agitation in the real part of the input impedance, the input tubes, and in the incoming radiation itself.

These three sources of noise are essentially of the same nature and can, therefore, all be combined in an equivalent noise power generator at the receiver input of internal power N_0P_0 in series with the received power P_r, i.e., with an internal impedance equal to the antenna impedance. N_0 is the receiver "noise factor."

$$P_0 = 1.6 \cdot 10^{-20} B \text{ watts}$$

where B is the effective noise bandwidth of the receiver before limiter or demodulator, and P_0 is the minimum possible noise power. The numerical coefficient is actually $4KT$ where K is Boltzmann's constant and the absolute temperature T is arbitrarily taken to be $T=293$ degrees (20 degrees centigrade).

The required power P_r can be expressed in terms of P_0 by what could be called a "system noise factor N" given by

$$\frac{P_r}{P_0} = N = \frac{N_0\, N_1\, N_2}{N_3}. \tag{7}$$

Equation (7) determines the power P_r if the various factors N_0, N_1, N_2, N_3, and the bandwidth B are known.

At microwave frequencies, the noise factor N_0 for good receivers is generally between the following limits: $10 < N_0 < 50$.

N_1 is the output signal-to-noise ratio. N_1 might vary from 20 decibels for teleprinters to 60 decibels for high-class music. For telephony, 40 to 50 decibels are generally considered adequate for good commercial quality.

N_2 is a factor which takes into account the presence of a certain number of receivers in tandem. Normally, the noise of each receiver will be added to the total and therefore N_2 is simply equal to the number of relays in series. N_2, resulting from accumulation of noise in repeaters, may equal zero decibels for all practical cases having any number of repeaters if pulse-code modulation were used.

N_3 is the improvement factor resulting from the type of modulation, bandwidth, type of limiters and demodulator, etc.

For frequency and pulse-time modulations, N_3 is proportional to the square of the ratio of the transmission bandwidth to the bandwidth covered by the signal at least so long as the ratio P_r/N_0P_0 is larger than 6 to 12 decibels to let the limiters operate correctly. The choice of the factor N_3 will determine the bandwidth and P_r or, inversely, the choice of the bandwidth will determine N_3 and P_r.

These calculations are straightforward for simplex systems. For multiplex systems, they become more involved because the signal-to-noise ratio per channel alone is determined, also because often the bandwidth is determined not only by the desired improvement factor but also by the amount of cross talk that can be tolerated. In all cases, it seems theoretically advantageous to use effectively as wide a bandwidth as possible without reaching a value where selective fading and amplification difficulties cause trouble. From this point of view, time-division methods with amplitude keying are attractive. Indeed, in this case, the power to be considered for operating the limiters should be the peak power. For a given number of channels, the peak power can be made proportional to the bandwidth keeping the average power constant. In other words, for frequency modulation, the average power necessary increases with bandwidth for a given input-to-noise ratio, but in pulse-amplitude keying, the average power necessary increases much more slowly. As a first approximation, it is constant; when the probability of noise peaks overriding the limiter level is taken into account, the average power should increase as the logarithm of bandwidth.

It is not intended to reproduce the calculations leading to possible optimum values of the various factors N, but only to indicate practical limits. A typical example follows:

$$\begin{aligned}
N_0 &= 15 \text{ decibels (average noise factor)}\\
N_1 &= 45 \text{ decibels (telephony)}\\
N_2 &= 10 \text{ decibels (10 repeaters)}\\
\underline{N_3} &= \underline{-20 \text{ decibels}} \text{ (improvement factor)}\\
N &= 50 \text{ decibels and } B = 8 \cdot 10^6 \text{ cycles}\\
P_r &= 1.6 \cdot 10^{20} \times 8 \cdot 10^6 \times 10^5 = 13 \cdot 10^{-9} \text{ watt.}
\end{aligned}$$

The necessary power to operate the receiver is of the order of several millimicrowatts. Knowing the power required at the receiver, the transmitted power will be determined by (1).

$$\begin{aligned}
A_0 &= 60 \text{ decibels (free-space attenuation)}\\
\underline{A_1\, A_2\, A_3} &= \underline{30 \text{ decibels}} \text{ (allowance for fading)}\\
A &= 90 \text{ decibels}\\
P_t &= 10^9 \times P_r = 13 \text{ watts.}
\end{aligned}$$

The transmitter power necessary to operate the link is of the order of several milliwatts if no allowance is made for fading and of the order of several watts with such an allowance.

The figures above are, of course, only representative and each case has to be calculated specially.

5. *System Engineering*

In the previous paragraphs, consideration has been given to the most important theoretical aspects of microwave relay links. Of practical importance are many other aspects, such as: type of tubes available, cost of towers, type of power supply available, cost of ground installations, roads, houses, and integration of the radio link into the existing telephone network.

Progress during the war in microwave tubes indicates that practically all types of tubes necessary for such links can be designed and some will soon be commercially available.

The magnetron has proved to be the most efficient microwave generator so far but it cannot be modulated easily. Probably, it will be possible to design frequency-modulated magnetrons and it is already possible to key magnetrons on and off as required for pulse-time modulation.

Velocity-modulated tubes of the reflex type are easily frequency modulated but have poor efficiency and, so far, are available at low-power levels only. The main tool lacking is a broad-band amplifier. Very promising results have been obtained during 1946 with models of traveling-wave amplifiers and of a broad-band velocity-modulated tube. As has been pointed out, many problems of radio relaying would disappear if it were possible to amplify at each relay without changing frequency.

The towers necessary for obtaining the desired antenna height represent a considerable part of the investment for radio relay links. The height and rigidity of the tower are probably the most important elements in determining the technical characteristics of the link.

It is quite obvious that from a maintenance point of view it is preferable to house all the equipment at ground level and install the antennas only at the top of the tower. But such a solution raises the difficult problem of transmission between the transmitters and receivers and the antennas. Transmission lines, either coaxial or wave guides, introduce additional losses and are susceptible to failure under unfavorable climatic conditions. One very attractive solution, possible only at the higher frequencies, is to install the antennas also at ground level and to use a mirror at the top of the tower.

An entirely different school of thought is that expensive towers are justified as a permanent investment. Even if the equipments should be modified or replaced by future developments, the towers could still be used.

Actual cost estimates for the 24-channel system outlined in the appendix, Section 7.2, show that a 100-foot steel tower designed to house all the equipment on top costs 20 percent more than the repeater equipment. This does not include any allowances for purchase of grounds or for eventual road construction.

For economic reasons, radio relay links in many cases should operate with unattended repeaters. It seems likely that the repeater equipment proper could be designed to operate without attention for several months but it is not yet clear how to design an independent power generator for such a long life. When a source of power is not available in the immediate vicinity of the relay station, it will be necessary to set up an independent power supply.

Power generators of 500 to 2000 watts, guaranteed for reliable continuous operation for several thousand hours, do not seem readily available. Radio relays, therefore, are often in the paradoxical position that, though they have eliminated physical connections between relay stations for carrying signals, it is necessary to resort to such connections for carrying power for operation of the equipment.

The integration of a radio relay link into an existing communications network is not essentially different from similar problems for carrier transmission links, but radio engineers often are not aware of the considerable amount of system planning required to set up or to expand the wire and cable networks.

Requirements for ringing, signaling, and dropping channels might react on the design of the radio link itself or at least on the terminals. Links designed to operate with frequency-division multiplex are generally connected to existing carrier terminals and the system planning is done by wire-transmission engineers.

Links designed to operate with time-division multiplex are generally connected to the switchboard itself and include the multiplexing terminals. A more intimate coordination between radio and telephone planning is, therefore, necessary. Time division here offers attractive possibilities such as direct-current signaling and quite simple methods for dropping channels.

6. *Conclusion*

An attempt has been made to give a general view of the various problems involved in the engineering of radio links with repeaters. As for most similar problems, there is no single answer and the final design is essentially a matter of compromises between conflicting requirements. Further experiments are required before the optimum compromises will be known.

Radio relay links definitely seem promising both on technical and economic grounds, but there is not enough operating experience available to indicate how reliable these new communication systems will be when compared to wire or cable links. The economic balance between radio and wire transmission will, indeed, depend essentially on the degree of reliability desired.

It seems unlikely that radio relay systems could be quite as reliable as buried cables but they might well in the near future be able to compete with open-wire links and they will, also, facilitate the introduction of new, broadband services.

7. *Appendixes*

7.1 PARIS-MONTMORENCY MICROWAVE LINK (EXPERIMENTAL EQUIPMENT)

A. *General System Characteristics*

Multiplexing: Frequency Selection, 12 Channels
Channel Modulation: Single-Sideband Amplitude Modulation
Carrier Modulation: Frequency Modulation with Negative Feedback
Number of Repeaters: None
Distance Between Terminals: 24 Kilometers (15 Miles)

B. *Main Technical Characteristics*

Frequencies: 3333 Megacycles One Way, 3000 Megacycles the Other Way
Transmitter Power: 30 Watts with Velocity-Modulated Output Amplifier Tube
Antenna Aperture: 0.5 Square Meter (Horn-Type Antenna)
Antenna Gain: 22 Decibels Referred to Half-Wave Dipole
Radiation Beam Width: 6 to 7 Degrees in Both Horizontal and Vertical Planes
Radiation Polarization: Horizontal at 3333 Megacycles, Vertical at 3000 Megacycles
Input Signal Bandwidth: 12 to 60 Kilocycles
Carrier-Frequency Deviation: ±360 Kilocycles
Nonlinear Distortion at the Transmitter: −65 Decibels
Over-All Output Signal-to-Noise Ratio for Each Channel: 55 Decibels
Over-All Output Signal-to-Cross-Talk Ratio for Each Channel: 65 Decibels.

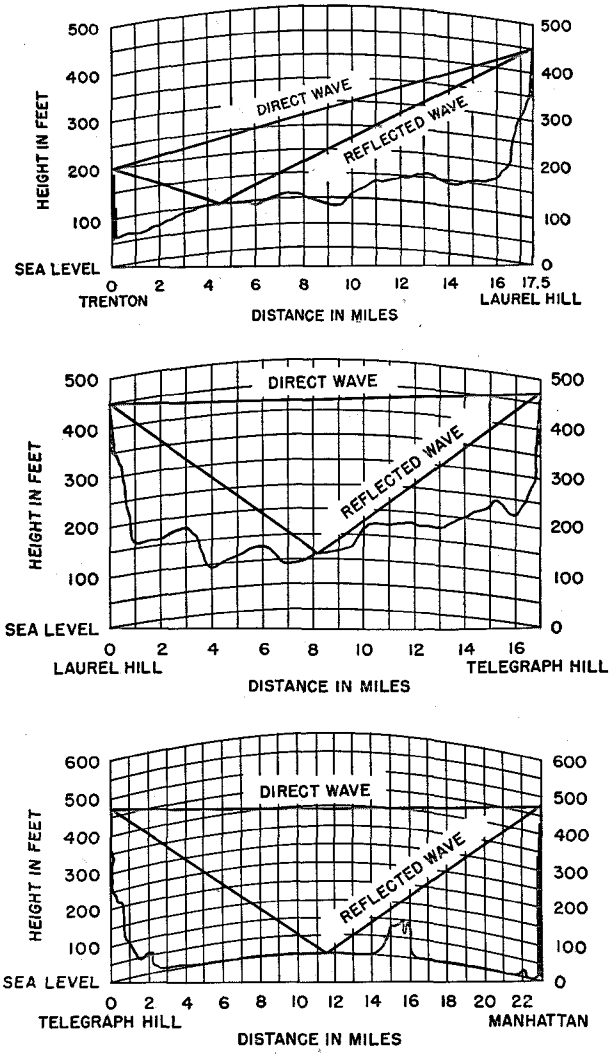

Fig. 3—Topographic paths of the three links in the circuit between the lower tip of Manhattan in New York City and Trenton, New Jersey.

7.2 New York-Trenton Microwave Link (Experimental Equipment)

A. General System Characteristics

Multiplexing: Time Division, 24 Channels
Channel Modulation: Time Modulation
Carrier Modulation: Amplitude Keying
Number of Repeaters: 2
Altitudes and Distances Between Stations: See Fig. 3
General Map of the Links: See Fig. 4

B. Main Technical Characteristics

Frequencies: 1375 to 1425 Megacycles
Transmitted Power: 2 Watts Average, 20 Watts Peak
Antennas: Paraboloid 10-Foot Diameter with Horizontal Dipole Radiator
Radiation Beam Width: 5 Degrees Vertically, 7 Degrees Horizontally
Receiver Bandwith: 8 Megacycles
Signal Bandwidth: 2.4 Megacycles at Video Frequencies
Signal-to-Noise Improvement Factor: 20 Decibels
Receiver Noise Factor: 15 Decibels
Over-All Frequency Response per Channel: Constant within ±2 Decibels from 200 to 3400 Cycles
Over-All Distortion in Each Channel: 2 Percent Harmonic Content for ±1-Microsecond Time Modulation
Over-All Signal-to-Noise Ratio in Each Channel: 45 Decibels ±1-Microsecond Time Modulation
Over-All Cross-Talk-to-Noise Ratio in Each Channel: 55 Decibels for ±1-Microsecond Time Modulation.

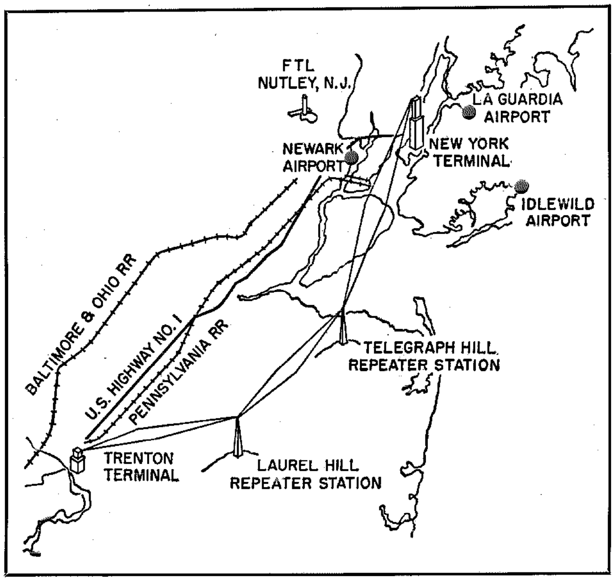

Fig. 4—Map showing locations of the four stations comprising the New York City-to-Trenton circuit.

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Pulse-Time-Modulated Multiplex Radio Relay System—Radio-Frequency Equipment

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TERMINAL EQUIPMENT for a radio relay system[1] has already been described. This paper covers the additional apparatus consisting of radio-frequency equipment, antenna and transmission-line systems, towers, power sources, and auxiliaries such as monitoring devices. Also included are test results on an experimental relay system incorporating the apparatus described.

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1. *Radio Relay Considerations*

1.1 Frequency of Operation

The choice of operating frequency is of primary consideration for radio relay links irrespective of transmission and modulation methods. From a theoretical viewpoint, any radio frequency consistent with bandwidth requirements may be utilized satisfactorily; from the practical point of view, the range of frequencies that can actually be utilized is limited.

Except for restricted applications in nonpopulated areas, the frequencies below 100 megacycles per second are unavailable because of the congestion of other services. For frequencies between 100 and 1000 megacycles, large transmitter powers must be used for a reasonable signal-to-noise ratio because of the limited size of practical antennas and the attendant small antenna gain. This limits the application to short links where considerable amounts of power are available, and makes impracticable the use of many repeaters over long distances where the provision of primary power sources is difficult.

Ignoring for the moment the intermediate microwave region, the application of frequencies above 7000 or 8000 megacycles to radio links is again limited. For these frequencies, reasonably sized antennas give large gains and make the transmitter power requirements small. Unfortunately, economical design limits the rigidity of a given tower used to meet the line-of-sight requirements. This in turn sets a lower limit to the beam angle utilized and, therefore, to the gain achievable. Unless automatic follow-up mechanisms are used with the antennas, practical beam widths are limited to approximately 1 or 2 degrees (3-decibel points) for reasonable tower rigidity. Otherwise, undue signal fluctuation may result from tower sway.

There are additional reasons for limiting the frequency to the values given. As frequency is increased, absorption and diffraction effects become more important. Power generation and reception difficulties, although of a temporary nature so far as future development is concerned, increase with frequency. Then, too, frequency stability, transmission lines, components, and other allied system items introduce difficulties.

This brief examination of the range of frequencies suitable for relay operation has been rather general but, if the items indicated are studied in detail, it will be found that the range from approximately 1000 to 7000 megacycles should offer the most likely prospects for such application, at least for the present time and with current techniques.

1.2 Repeater

A second problem of major magnitude for long radio relays is the lack of a true repeater-amplifier for the frequencies that can be used. For long relays and especially for multiplex operation, severe requirements are put on the linearity of the repeaters to prevent undue distortion and cross talk. The invention of negative feedback for low-frequency narrow-band amplifiers has solved the repeater problem for that service and for cable application. A similar solution has not been successfully applied to the higher frequencies and hence other means of amplification must be used.

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[1] D. D. Grieg and A. M. Levine, "Pulse-Time-Modulated Multiplex Radio Relay System—Terminal Equipment," *Electrical Communication*, v. 23, pp. 159–178; June, 1946.

Lacking a direct amplifier for wide-band high-frequency work, indirect amplification is utilized. Thus, the high frequency as received is transposed to an intermediate frequency. All amplification takes place at this intermediate frequency, and the amplified signal is used to remodulate a high-frequency carrier.

Unfortunately, the mechanics of this transposition inherently introduce distortion which rules out amplitude modulation for any reasonable number of repeaters. Even for the constant-carrier type of frequency modulation where amplitude nonlinearities are of minor importance, phase nonlinearities introduced by intermediate-frequency amplification must be carefully controlled.

1.3 Frequency Stability

An important item in a radio relay is the stability of the component transmitters and receivers included in the system. This stability is required not only for the conservation of frequency spectrum but is important because minimizing bandwidth also reduces the total noise entering the system. If a direct amplifier could be utilized, the over-all frequency stability of the system would be determined by the terminal transmitters and receivers. Suitable precautions, even though necessitating relatively complex equipment, could then be taken at these single points. Where indirect amplification involving change of frequency is utilized, the frequency spectrum required for the over-all link is increased by the summation of the percentage instability of the various receivers and transmitters comprising the link. Special precautions must, therefore, be taken to maintain frequency stability.

1.4 Transmission Lines

A problem of considerable importance particularly at the repeater points is the method of feeding the antennas. The radio-frequency equipment may, of course, be placed close to the antennas on top of the towers with only a short transmission line connecting the units. This, however, requires a tower structure of considerable strength and complexity, capable of safely supporting all this equipment. Alternatively, the radio-frequency equipment can be located at the base of the tower and the antenna connected through a transmission line equal in length to the tower height.

At the lower frequencies, a solid-dielectric transmission line can be used with some loss of power. With higher frequencies, however, the loss in the solid-dielectric line becomes prohibitive, and either a gas-filled line or wave guide must be used. With these two latter types, considerable difficulties are encountered in, among other things, maintaining airtight seals, providing for expansion, and preventing vapor condensation.

A partial solution to the problem is to divide the equipment so that the video-frequency apparatus is at the base of the tower and only the radio-frequency equipment required for transmission is installed at the tower top. The connection between the two units is then made by video-frequency cable.

This method is open to serious criticism from the maintenance point of view as it is necessary to mount the tower to service or adjust the radio-frequency equipment. A possible solution, which has been partially investigated, is the use of radio-frequency mirrors on the top of the tower with the radio equipment located at the base. A directive antenna concentrates the beam on the mirror which is arranged at the proper angle to obtain effective transmission. This method increases the transmission path by a negligible amount equal to the height of the tower, but effectively eliminates the transmission line.

1.5 Power for Repeaters

An additional problem is that of a primary power supply for unattended repeaters. If commercial electric power is available at the repeater site, no undue problems are encountered other than that of forestalling interruption of service. If, however, such a repeater is located remotely from available power sources, which in general would be the case where the system runs through undeveloped territory or takes advantage of isolated high points of ground, a serious problem is faced. Unlike cable operation where the power can be supplied over the same medium as used for connecting the repeaters, local power must be provided. This dictates the use of a local fuel dump if natural sources such as wind or water power are not available, together with the necessary electricity-generating equipment. If the repeater is relatively inaccessible, the power source must be capable of supplying power for considerable periods of time without attendance. This in turn necessitates efficient and reliable means for generating electric power and also indicates the desirability of minimizing power consumption.

1.6 Two-Way Operation

In most cases of relay operation, particularly for telephone service, the link must be capable of two-way performance. This type of operation can, of course, be achieved by installing repeater equipment for each direction of transmission at the common repeater sites. If continuous-wave transmission is used, an alternation of frequencies is necessary to prevent feedback between transmitter and receiver. This puts a severe requirement on the selectivity and image characteristics of the repeaters, particularly if the total frequency band is to be minimized.

It is theoretically possible to use only two frequencies by proper alternation in each direction. In practical installations, however, the topography may be such that repeater hopping or by-passing of the immediately following repeater may take place. This may be avoided by additional frequency transpositions. In addition, two-frequency operation calls for a high discrimination ratio between the directional antennas at the common repeater site and this also may be impossible because of local topography. If the attempt is made to combine the four antennas required for normal two-way operation into two antennas, one for each direction of transmitting and receiving, an extremely difficult design problem is raised unless a plurality of frequencies are used with highly selective repeaters.

1.7 Reliability and Maintenance

A most important item is the reliability and maintenance requirements at the repeater. For a long link containing many repeater elements, the performance of the poorest repeater determines the over-all operation of the entire link so far as reliability is concerned. When it is considered that a single repeater may contain as many as twenty vacuum tubes, a long link would have the equivalent of many hundreds or thousands of tubes in series. Failure of any one of these tubes may render the link inoperative unless suitable precautions are taken.

Reasoning similarly, it is evident that the over-all stability of the repeaters must be largely independent of weather conditions, primary-power fluctuations, tube aging, and similar factors. Where the repeater equipment is accessible, the degree of adjustment and servicing allowable may involve daily or weekly checking consistent with the economies of operating the relay link. If the repeater is located remotely, the requirements of maintenance are much more severe and it may be necessary that the equipment operate for several months completely unattended in all respects.

To meet the severe requirements of reliability, it is necessary to operate most of the equipment at low rating and to provide duplicate apparatus with automatic switch-over mechanisms. A remote telemetering arrangement should be provided to indicate at the terminals of the link the operating condition of the various equipments.

1.8 Applicability of Time-Division Multiplex

The use of time-division multiplex together with pulse modulation, such as pulse-time modulation,[2] solves many of the radio-frequency transmission problems previously mentioned.

An inherent characteristic of this system of modulation is that only one increment of the signal is transmitted at any one instant of time and that incremental signal is characterized by the relative timing of the transmitted pulse. This on-off type of operation, unlike straight amplitude or frequency modulation, may utilize a considerable number of methods of radio-frequency generation and hence the type of modulation need not be considered in the choice of operating frequency.

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[2] E. M. Deloraine and E. Labin, "Pulse Time Modulation," *Electrical Communication*, v. 22, n. 2, pp. 91–98; 1944.

A most important aspect is that no matter what distortion is produced in the amplifying means, only a minimum of cross talk or distortion of the modulating signal can take place if a sufficient bandwidth is provided to prevent carry-over from one pulse to another. Hence the repeater problem is considerably simplified and reduced to the design of a repeater amplifier of the proper noise characteristics, gain, bandwidth, and efficiency.

Additional characteristics of pulse transmission include the possibility of transmission and reception on a common radio frequency similar to radar operation with a consequent reduction in bandwidth as well as a simplification in equipment requirements. Indirectly, a gain in over-all frequency stability may be achieved also as fewer frequency changes are necessary. A further characteristic of importance is the ease of drop-channel operation at branching repeaters which allows any number of channels to be dropped and new channels to be inserted without necessitating the handling of large groups of channels. Additional advantages include signal-noise improvement possibilities of the subcarrier modulation which can be further enhanced by the signal-noise improvement achieved by the over-all radio-frequency transmission method. Further, constant-level transmission allows the output signal level to be substantially independent of propagation vagaries.

2. *Radio-Frequency Equipment*

2.1 GENERAL

The design of a radio link with a specific set of operating characteristics requires a knowledge of several transmission factors. These factors include over-all signal-to-noise ratio, repeater gain and power, number of repeaters, antenna size and beam angle, propagation variations such as fading, and height of supporting structures necessary to yield the requisite propagation characteristics.

The largest proportion of noise in the link originates at the receiver input. For each receiver, the theoretical thermal agitation noise[3, 4, 5] (Johnson noise) at the antenna must be multiplied by the noise factor of the receiver. This factor takes into account the increase in the receiver noise resulting from such component parts as the mixer, local oscillator, and first intermediate-frequency stage. Assuming linear addition of noise power, the over-all link noise is the total receiver noise multiplied by the number of link hops.

The signal power received at the terminals of the link is determined by the output signal-to-noise ratio desired. For an amplitude-modulated system, the required received power is simply the total link noise multiplied by the desired signal-to-noise ratio. For systems where the signal-to-noise improvement over amplitude modulation is obtained by a specific type of transmission, this latter power is divided by the improvement factor. For the time-modulated relay to be described, the improvement factor is of the order of 16 decibels and thus the required power is reduced by this amount. Knowing the transmission attenuation factor, which is determined by the spacing between repeaters and the antenna gain, the repeater gain and power can be computed.

The repeater spacing, which is an important consideration, is determined by economic factors as well as technical requirements. The greater the number of repeaters, the less gain required of each repeater and the lower the antenna height may be. However, an increase in the number of repeaters increases the amount of apparatus required and the cost. A study of the various

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[3] D. O. North, "Absolute Sensitivity of Radio Receivers," *R.C.A. Review*, v. 6, January, 1942.

[4] E. W. Herold and L. Malter, "Some Aspects of Radio Reception at Ultra-High Frequencies," *Proceedings of the I.R.E.*

Part I—"The Antenna and the Receiver Input Circuits," v. 31, pp. 423–438; August, 1943.

Part II—"Admittances and Fluctuation Noise of Tubes and Circuits," v. 31, pp. 491–500; September, 1943.

Part III—"The Signal-to-Noise Ratio of Radio Receivers," v. 31, pp. 501–509; September, 1943.

Part IV—"General Superheterodyne Considerations at Ultra-High Frequencies," v. 31, pp. 567–574; October, 1943.

[5] H. T. Friis, "Noise Figures of Radio Receivers," *Proceedings of the I.R.E.*, v. 32, pp. 419–422; July, 1944.

factors involved has indicated[6] that a proper balance is reached for spacings of the order of 20 to 30 miles unless special topographical conditions prevail.

As directional transmission is desired, the type of antenna utilized concentrates the transmitted energy in a narrow beam, and hence the required repeater power is reduced by the factor of antenna gain at both the receiving and transmitting points. As has been indicated previously, the beam angle is limited by mechanical considerations of tower rigidity for the higher frequencies. For the particular frequencies used in the present relay, the limit of beam angle is not reached, but a mechanical limitation is imposed by the size of reflector dish that can be accommodated on top of the tower.

The prevalence of fades and other propagation anomalies must be provided for in the determination of the transmitter power. These variations, in general, will affect only the final signal-to-noise ratio in the link for pulse-time modulation rather than the level of the signal transmitted. Where a large number of repeaters are used, extending over a large geographical area, it can be assumed that the fading conditions do not affect all repeaters simultaneously and hence the fading factor need not correspond to the sum of the individual factors.

Where the higher frequencies are used, the height of the antenna is determined not only by the condition of line of sight, but must include an additional height so that the summation of the free-space wave and the main reflected wave yields a wave at least equal in magnitude to that of free-space propagation. Normally, advantage is taken of topography to provide line-of-sight conditions and the tower need only be tall enough to clear local obstructions such as trees.

The radio-frequency equipment has been designed to operate in an experimental pulse-time-modulated multiplex radio link having the overall characteristics specified in reference 1. As this equipment was designed for experimental purposes only, it was not necessary to include all of the factors that would be obligatory in a commercial installation. However, provisions were made for testing facilities to duplicate conditions representative of commercial operation. This link equipment comprises terminal radio-frequency apparatus as well as a repeater. The terminal equipment transposes the outgoing time-modulated pulse series to radio frequencies, and in addition translates the incoming radio-frequency pulses back to pulse frequencies for application to the multiplex terminal. The repeaters, of course, compensate for the attenuation along the radio path.

Fig. 1 illustrates the disposition of this equipment in a radio relay system. The link incorporates a transmitter and receiver at each radio-

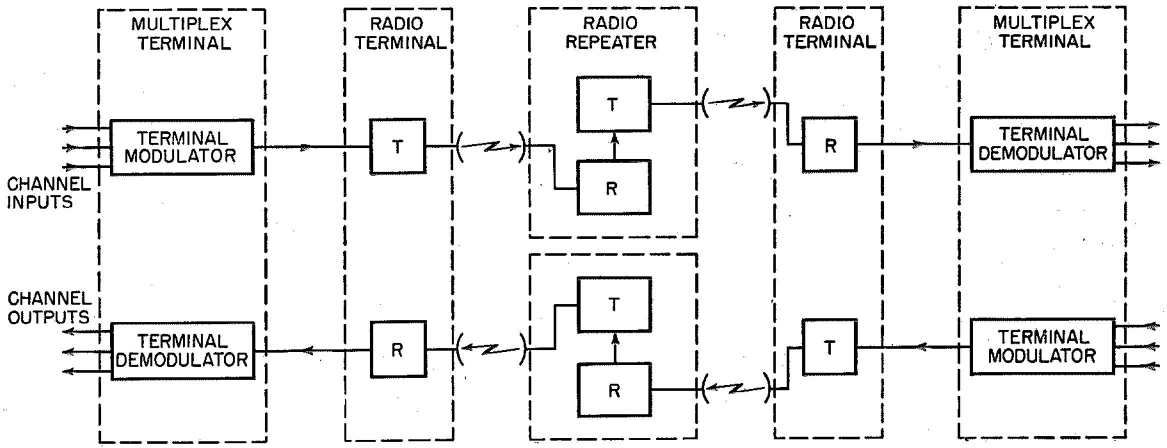

Fig. 1—Disposition of equipment in a radio relay system. *R* = receiver and *T* = transmitter. Only one repeater is shown; additional repeaters may be included.

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[6] C. W. Hansell, "Radio-Relay-Systems Development by Radio Corporation of America," *Proceedings of the I.R.E.*, v. 33, pp. 156–168; March, 1945.

frequency terminal and two receivers and transmitters, one for each direction, at the repeater stations. Except for minor differences in impedance termination at the receiver output and transmitter modulator input, the electrical and physical requirements for the radio-frequency terminals and repeaters are the same, and essentially identical equipment can be used for both applications. The description can therefore be limited to a receiver and transmitter which can be combined in the required manner to serve as either terminal or repeater equipment.

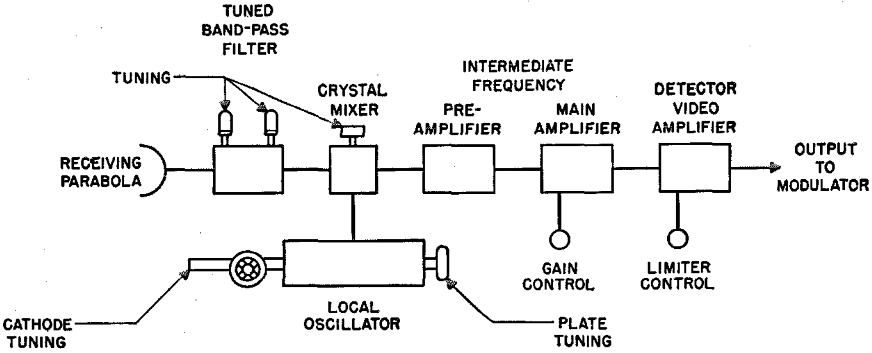


Fig. 2—The receiver consists of the components indicated in this block diagram.

2.2 Receiver

Fig. 2 shows the over-all receiver arrangement while Fig. 3 details the input end of the receiving circuit. The band-pass filter utilized to reduce interchannel interference consists of two micrometer-tuned coaxial cavities over-coupled to produce an 8-megacycle bandwidth having an attenuation of 50 decibels at the image frequency, 60 megacycles from resonance. The crystal mixer is a single tuned cavity connected to the output of the filter. The filter output, local oscillator, and crystal detector are coupled to the cavity by current loops along the cavity wall. The loops are adjusted for proper coupling and loading to produce the required bandwidth. An additional 22-decibel image-frequency attenuation is achieved by the tuned mixer, giving an over-all receiver image-rejection ratio of 72 decibels. Fig. 3 shows the crystal detector coupled to the first intermediate-frequency stage through a series tuned circuit made up of a tuned inductance and the grid-circuit capacitance to ground. This first stage is followed by two double-tuned stages. Each tuned circuit is shunted by resistance to provide the proper bandwidth. The resultant pass band, flat-topped and steep-skirted, is 8 megacycles wide at the 3-decibel points. The output stage is a triode cathode-follower which is connected by a 70-ohm cable to the intermediate-video-frequency chassis. The local oscillator is similar to the transmitter oscillator and will be described under that heading. Table I contains the technical details and specifications for the receiver. Front-panel metering jacks are provided for measuring

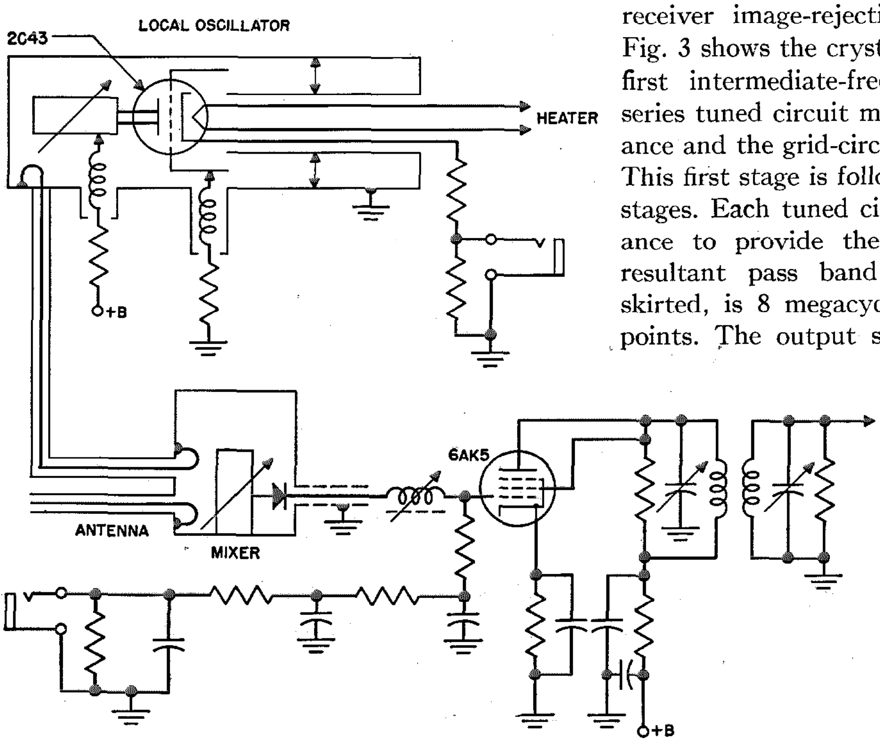


Fig. 3—Input end of the receiver showing local oscillator, input tuning, mixer, and first stage of intermediate-frequency preamplifier.

TABLE I

Characteristics of Transmitter and Receiver

RADIO-FREQUENCY TERMINAL-REPEATER	
Repeater (1 way)	
Transmitter	1
Receiver	1
Radio-Frequency Terminal	
Transmitter	1
Receiver	1
Transmitter-Receiver	
Size in inches	72×19×17
Weight in pounds	280
Power consumption in watts	280
TRANSMITTER	
Oscillator	
Tuning range in megacycles	1225 to 1325
Power output in watts, average	2.0
Bandwidth of pulse envelope in megacycles	5.6
Frequency stability in megacycles from −20 to +40 degrees centigrade	±0.5
Duty cycle	1:13
Oscillator tube	2C43
Oscillator circuit	Reflex coaxial cavity
Output impedance in ohms	50
Modulation	Grid pulsed
Pulse-time-modulation noise	Negligible
Cross talk	Negligible
Efficiency in percent	12 to 15
Modulator	
Circuit	Wide-band amplifier
Bandwidth in kilocycles	0.5 to 2800
Pulses per second	200,000
Pulse shape	Trapezoidal
Pulse width in microseconds	0.5
Pulse build-up time in microseconds	0.15
Pulse decay time in microseconds	0.15
Input impedance in ohms, adjustable	70 to 1000
Input level in peak positive volts	5
Output impedance in ohms	70
Output level in volts	60
Noise contributed to channel pulses	Negligible
Cross talk contributed to channel pulses	Negligible
Tubes	
Clipper (1)	6AC7
Driver (1)	6AG7
Cathode follower (2)	807
RECEIVER	
Input Section	
Frequency range in megacycles	1225 to 1325
Bandwidth in megacycles at 3-decibel points	8
Noise factor in decibels, approximate	14
Image rejection with input filter in decibels	72
Gain in decibels	20
Input impedance in ohms	50
Output impedance in ohms	70
Tubes	
Crystal mixer (1)	1N23B
1st radio-frequency amplifier (1)	6AK5
2nd radio-frequency amplifier (1)	6AG5
Output cathode-follower (1)	6J4
Local oscillator (1)	2C43
Intermediate- and Video-Frequency Amplifiers and Detector	
Input impedance in ohms	70
Bandwidth in megacycles at 1-decibel points	8

TABLE I—(*Continued*)

Noise factor in decibels	Less than 2
Gain in decibels	60 to 80
Minimum signal level in microvolts	200
Output volts, positive	5
Output impedance in ohms	70
Noise contributed to input signal	Negligible
Cross talk contributed to channel pulses	Negligible
Tubes	
Intermediate-frequency amplifier (4)	6AG5
Video-frequency detector (1)	6AG5
Video-frequency amplifier (1)	6AG5
Cathode-follower output (2)	6J4

the crystal-mixer current and oscillator grid and cathode currents.

The intermediate-frequency amplifier contains four amplifying stages, detector, pulse-frequency amplifier, and cathode-follower output stage. A study of the comparative gains of different interstage coupling networks for a wide-band intermediate-frequency amplifier resulted in the choice of triple-tuned circuits. Using 6AG5 tubes, a gain of 20 decibels per stage is realized for a bandwidth of 8 megacycles between 1-decibel peaks. Fig. 4 shows the over-all selectivity curve of the receiver including the selectivity of the intermediate-frequency amplifier and preamplifier.

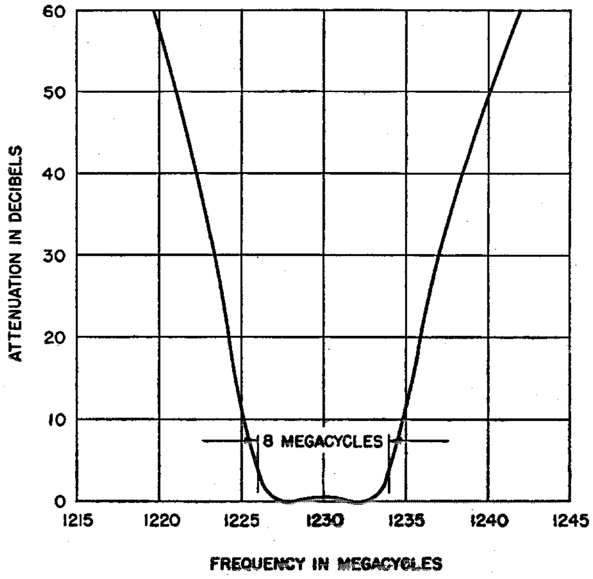


Fig. 4—Over-all selectivity curve of the receiver. A bandwidth of 8 megacycles is obtained at points corresponding to an attenuation of less than 3 decibels. This characteristic results from the combined selectivities of the preamplifier and the main intermediate-frequency amplifier, each of which has an 8-megacycle bandwidth at approximately 1-decibel attenuation points.

The detector and pulse amplifier are operated as limiters to remove noise from the top and bottom of the pulses and to make the output voltage essentially independent of variations in the received signal.

Two controls are provided in the intermediate-frequency unit, a gain control in the first stage and a bias control in the detector. In operation, these controls are set for optimum signal-to-noise ratio over a normal range of received signal-strength variations.

2.3 TRANSMITTER

Fig. 5 illustrates the transmitter, consisting of a modulator and radio-frequency oscillator. The transmitter utilized in the repeater differs from the terminal transmitter in that a delay line is interposed between the modulator and radio-frequency oscillator to transmit the pulse at a later time than it is received. This, of course, reduces the possibility of "ringing" between the transmitter and receiver and permits their frequencies to be closer together.

The modulator accommodates the proper frequency band as determined by the allowable carry-over between pulses. With the present design, all circuits are capable of amplifying up to 3.5 megacycles although the frequency spectrum of the transmitted pulses is of the order of 2.8 megacycles. The low-frequency design was considerably simplified by the absence of appreciable components below the pulse base repetition rate. The low-frequency cutoff is of the order of 500 cycles. The various amplifier stages clip both the top and bottom of the pulse, and the output pulse voltage is thus independent of variations of input level to the modulator.

Figs. 6 and 7 illustrate the oscillator design. This unit comprises a tuned coaxial grid-pulsed oscillator with a "grid-bell" type of re-entrant cavity utilizing the 2C43 lighthouse tube. The cathode line is tuned to either ¼ or ¾ wavelength by a piston. The plate is an open-ended ½-wavelength line adjusted by a screw-feed arrangement. The cathode resistor and its bypass capacitor provide a cut-off bias between

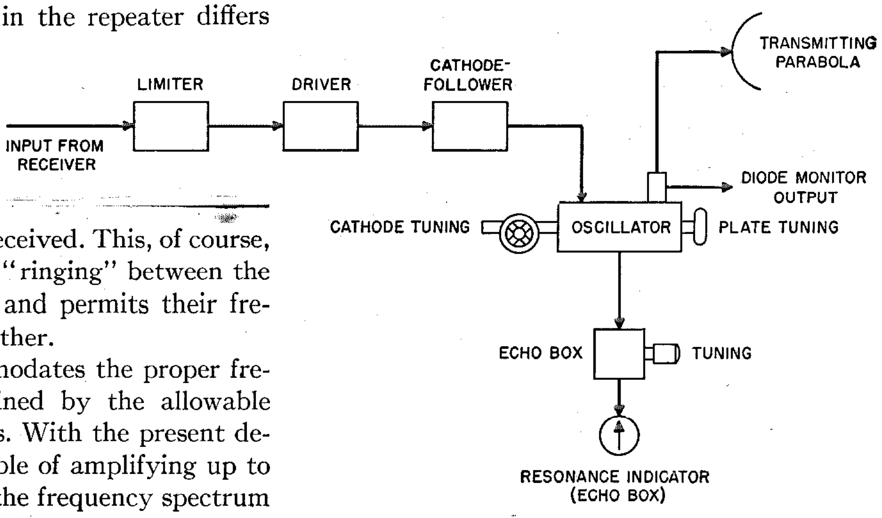


Fig. 5—Block diagram of transmitter. The modulation is applied to the oscillator through the cathode-follower. The tunable cavity resonator (echo box) and resonance indicator are used to monitor the output radio frequency.

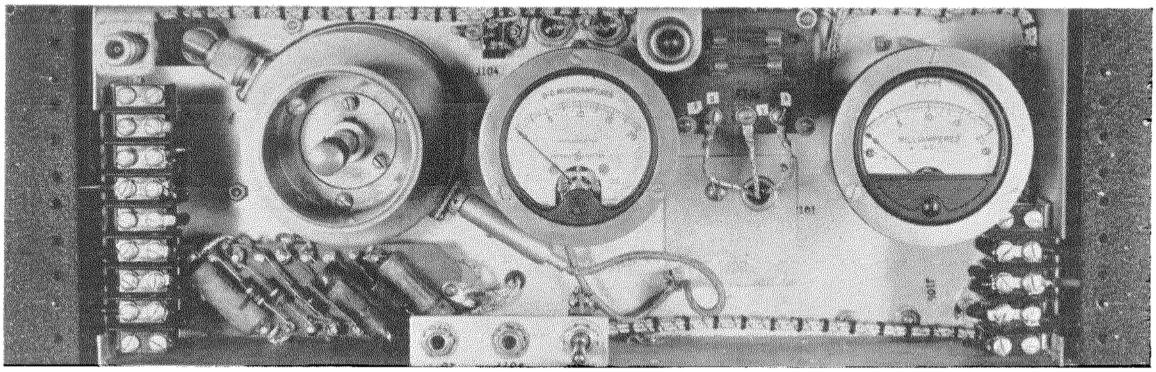


Fig. 6—Front view of transmitter. The echo box and its resonance-indicating meter are on the left end of the unit.

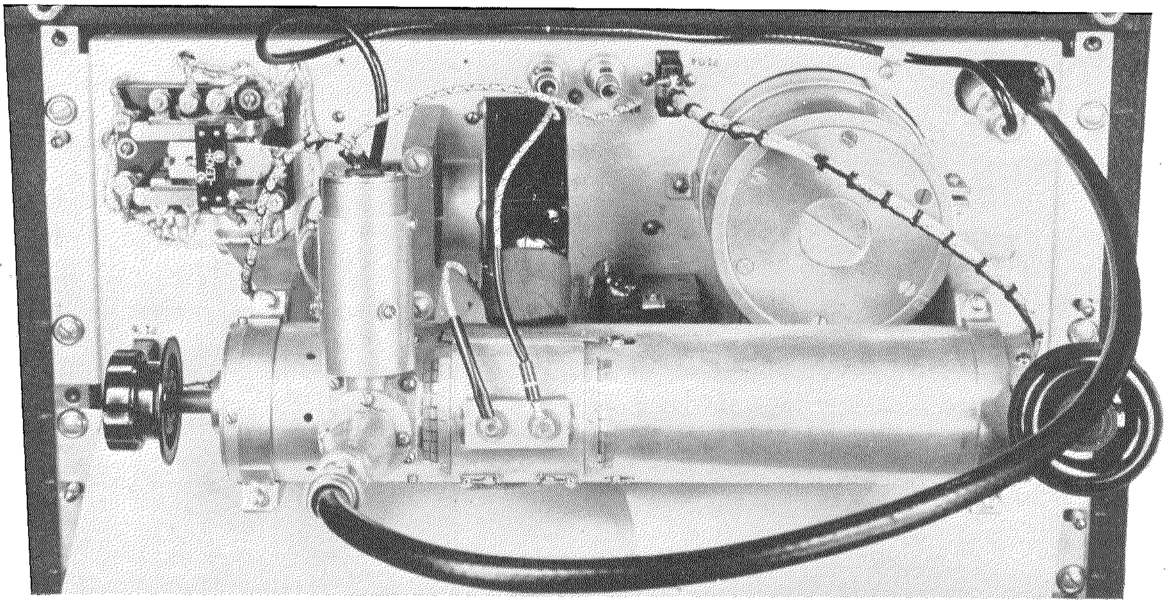


Fig. 7—Rear view of transmitter. The tuned coaxial oscillating circuit is at the bottom. The screw-feed adjustment at the left is for plate-circuit tuning. The cathode tuning control is at the right.

pulses. In addition, a grid resistor supplies the dynamic bias during oscillation.

A tunable echo box is loosely coupled to the oscillator cavity, and a crystal detector, also coupled to this cavity, indicates resonance. A linear diode monitor is coupled to the output line to provide an indication of the transmitted radio-frequency envelope. Temperature compensation of the oscillator is achieved by mechanical design, proper ventilation, and the avoidance of mechanical discontinuities in the resonant cavities. Fig. 8 shows the frequency drift of a unit for a temperature excursion from −38 to +60 degrees centigrade. The transmitter spectrum measured during pulse modulation is indicated in Fig. 9.

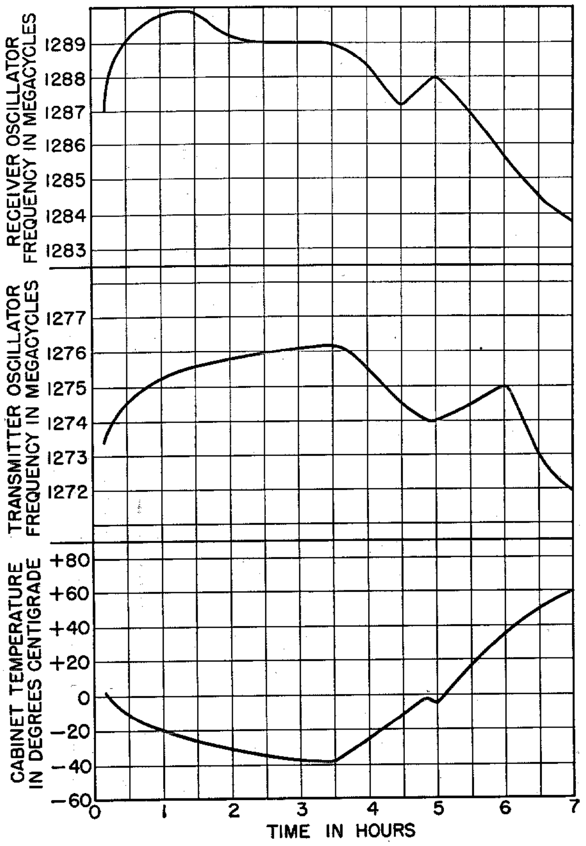


Fig. 8—Frequency variation of transmitting and receiving oscillators during 7 hours in which the temperature was changed from 0 to −38 to +60 degrees centigrade.

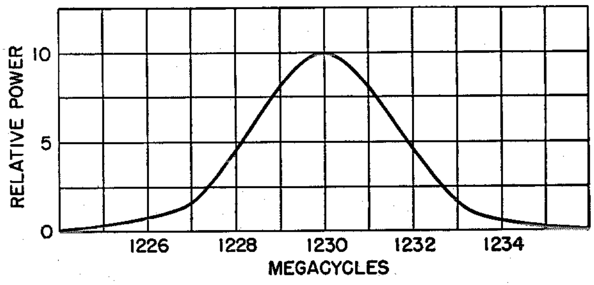
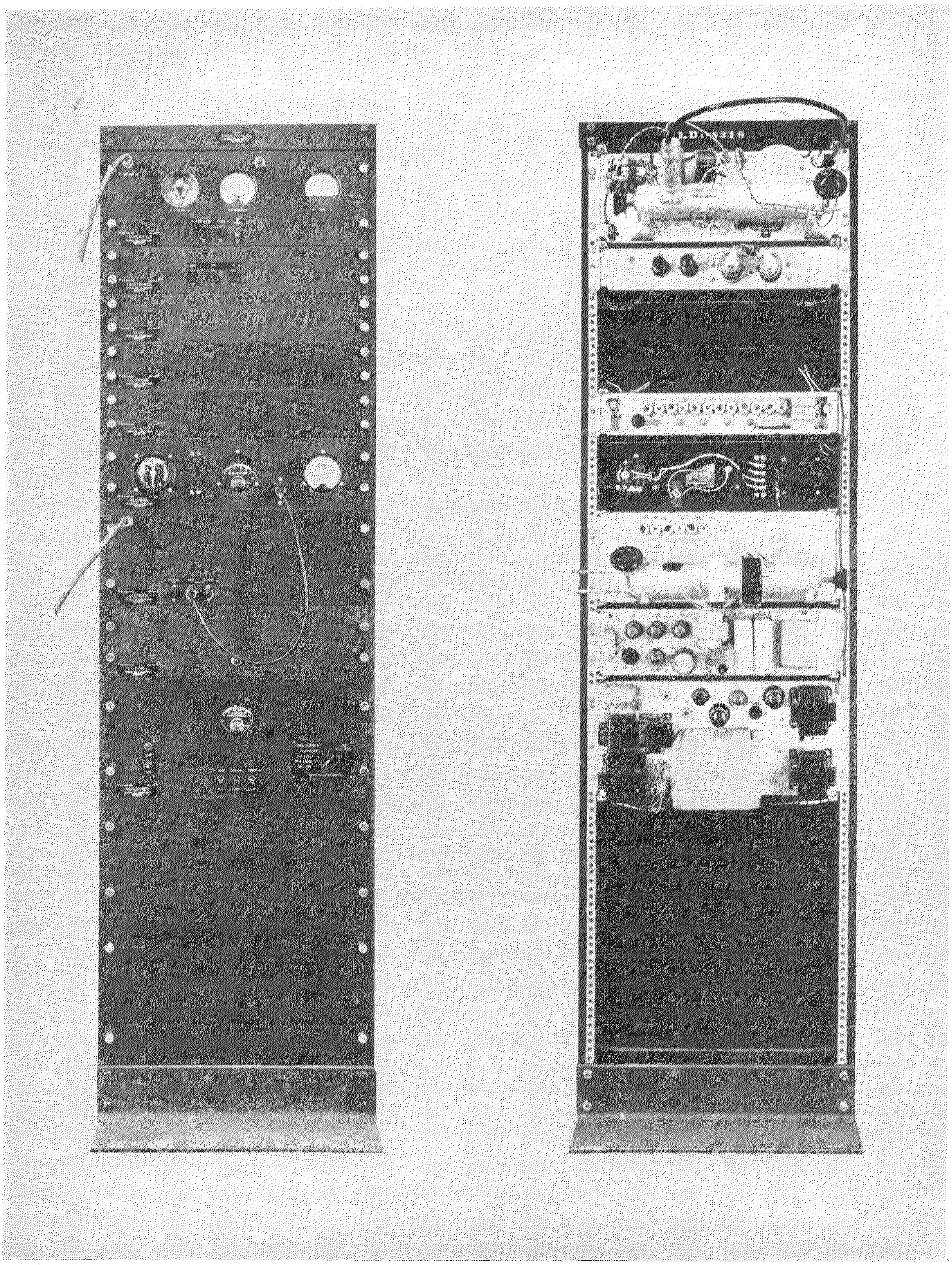


Fig. 9—Frequency spectrum of transmitter during pulse modulation.



Figs. 10 and 11—Front and rear views of radio repeater rack in which both the receiver and transmitter assemblies are mounted.

Table I indicates a few of the technical characteristics of the circuit. The identical type of oscillator, but without the echo box and monitor assembly, is used as the local oscillator in the receiver. Figs. 10 and 11 illustrate both the transmitter and receiver assemblies including power supplies and monitoring facilities mounted in a standard relay rack for repeater and terminal installation.

3. Relay Plant

3.1 GENERAL

Associated with the repeater elements of the relay are the important allied apparatus and structures which compose the "plant" of such a system. These items include:

A. Primary power source.
B. Antenna for radiating the radio-frequency energy and its associated transmission lines.
C. Towers for supporting the antenna at the required height.
D. Housing to protect repeater and other equipments from climatic conditions.
E. Miscellaneous mechanical and electrical items such as switch-over equipment, repeater monitors, etc.

The items enumerated, of course, are not necessarily limited to a single system of relay transmission such as pulse-time multiplex, but apply equally to all other types.

3.2 POWER SUPPLIES

A most important function of the auxiliary equipment at either a terminal or a repeater station is that of furnishing power to the radio-frequency units. This is further complicated by the varying conditions of operation.

The radio-frequency terminal station would normally be at a location where commercial electric power is available. In such a case, the principle problem is to forestall interruption of service resulting from failure of the commercial source. Several schemes may be used and choice depends on the degree of protection required. Where an interruption of a few minutes can be tolerated, an automatic engine-generator, electrically started on failure of the line power, may be used. Such a machine could carry the load only after a warm-up period of approximately one minute and would then be able to maintain service for an indefinite period, determined by the quantity of fuel available and the operating condition of the equipment.

If no interruption can be tolerated, it would be desirable to operate from a storage battery with a line-operated charger maintaining full charge. For further reliability this would be backed up by a gas-engine-generator automatically operated if the line power fails.

If there is primary electric power available at the repeater station, similar considerations apply. However, the power-supply problem is likely to be more severe at such a station for the site will probably be remote from normal sources of power. A significant characteristic of a repeater site is its elevation above the surroundings and this makes wind power entirely feasible. Fig. 12 indicates the practicability of wind-power sources in the United States. At a wind-powered station, the equipment can operate from a storage battery

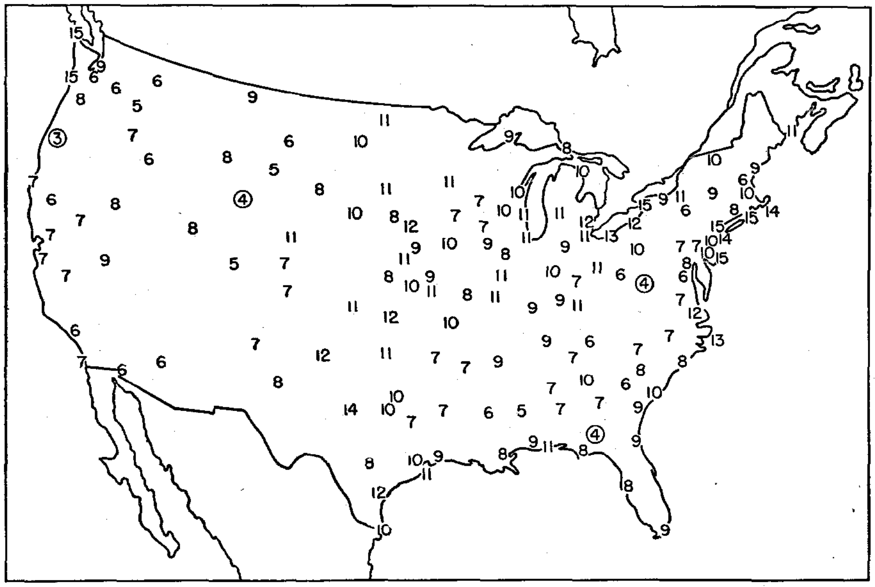


Fig. 12—Yearly average wind velocity in miles per hour for the United States of America. It is evident that adequate wind velocities are available, particularly on the high ground that would be chosen as relay sites, to make this an attractive source of power. The few circled values are considered to be inadequate.

of rather large capacity which is maintained charged by a wind-driven generator under both current and voltage regulation. To carry over periods of calm, an automatic gas-engine-generator is arranged to operate at any time the battery drops in charge.

To obtain complete reliability, a duplicate power source at each repeater station can be considered; for example, an independent power plant for each direction of operation. In the event of failure of one plant, the whole station load may be carried by the remaining unit. An experimental installation similar to that described has been set up and is illustrated in Fig. 13.

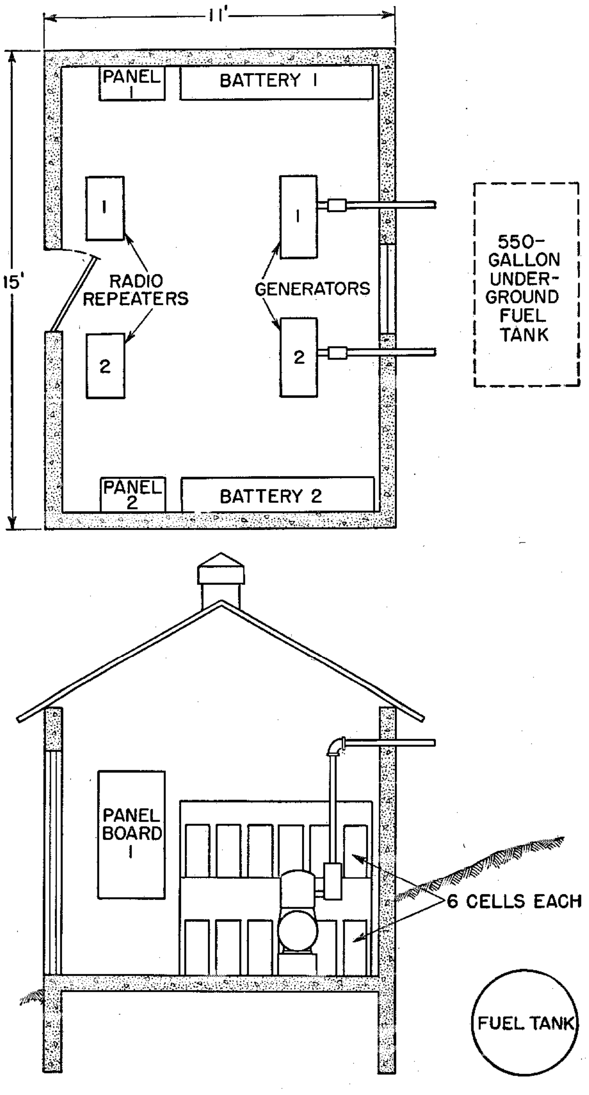


Fig. 13—Power plant for unattended repeater station.

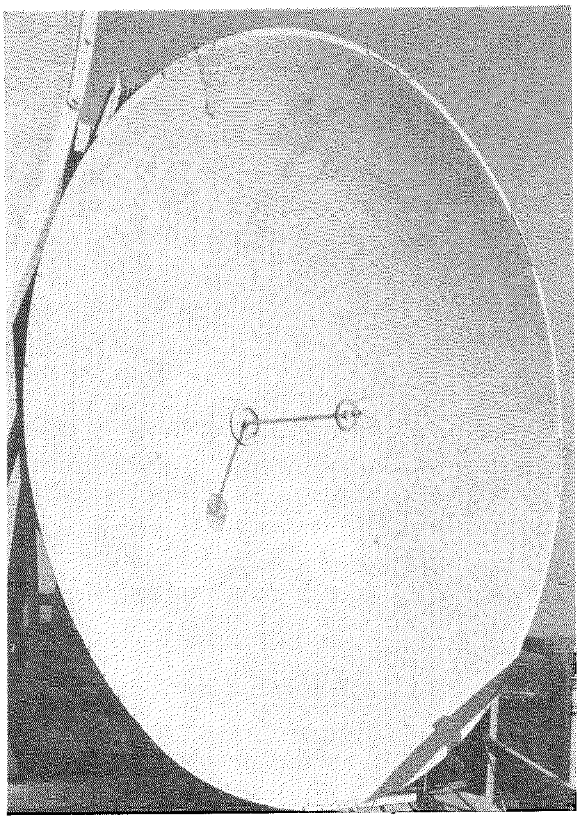


Fig. 14—Typical dipole and parabolic reflector for use in the 1000- to 3000-megacycle region. The enclosure around the dipole protects it from rain, snow, and sleet.

The equipment includes an automatic clock control which operates the engine at regular intervals daily to maintain the engine in running condition. Such an installation is capable of unattended operation over a period of many weeks, or even months, as it is necessary to service the batteries at intervals of only three to four months. The engine generally requires attention only after 200 or more hours of actual running time. Of course, criticism can be directed at the relative complexity of the installation. Unfortunately, reliable operation can be expected only through the use of several power sources, each backing up the other. Attempts are being made to simplify the problem by utilizing a heat engine or turbine. Also, the use of an atomic energy source is not ruled out for future application.

3.3 ANTENNAS

To obtain satisfactory range with reasonable power, a directional antenna is used. For fre-

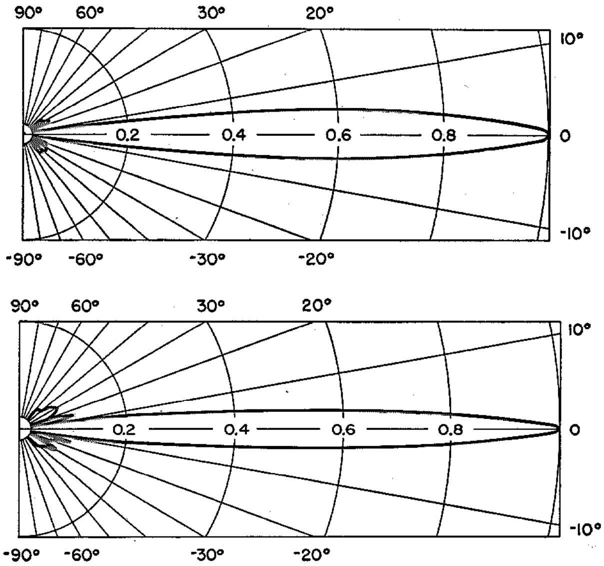


Fig. 15—Horizontal and vertical directivities of a 1400-megacycle (21.4-centimeter) dipole in a 10-foot parabolic reflector. Relative voltage output is plotted against angle of radiation. At the top, the directivity is measured with the dipole horizontal, while below it is vertical. Half power (3 decibels down) is equivalent to 0.7 relative voltage.

quencies between 1000 and 3000 megacycles, this generally consists of an enclosed dipole radiator with a parabolic reflector. At the higher frequencies, a similar reflector is used with a wave guide. Of course, other types of antennas and reflector systems may be used.

The reflector may be of perforated sheet or expanded metal. The expanded metal allows somewhat less wind loading under nonicing conditions and slightly less weight. A typical antenna is illustrated in Figs. 14 and 15. This antenna has been designed for an operating frequency of 1200 to 1400 megacycles with a gain of approximately 29 decibels over a half-wave dipole. For a two-way repeater station, four antennas are generally required unless special systems are used to combine the receiving and transmitting functions. In an actual installation, each antenna mount is generally adjustable over a considerable angular range, both horizontally and vertically, to allow accurate positioning.

3.4 Towers

For a working range of 30 miles between stations, a compromise tower height which can be utilized for most applications is of the order of 100 feet. This assumes that reasonable advantage is taken of topographical conditions. Under special conditions, this height may vary considerably. At sites having adequate elevation, the main requirement is that sufficient clearance be maintained with respect to immediate local obstructions such as trees and buildings.

The tower must be strong enough to withstand any possible combination of wind and ice loading which might occur in the area. If this condition is met, the structure will usually be sufficiently rigid to maintain antenna position with necessary constancy as well as providing support for additional equipment such as wind generators.

Relay towers may be constructed of steel, wood, or concrete. The steel structure is suitable for use most anywhere and is immune to severe fire, such as may occur in a forested area. The design and manufacture of a tower of this material can be controlled rather closely giving a product of known strength. Also, it is readily adaptable to quantity production, and maintenance requires little more than periodic painting and inspection. The chief disadvantage is the cost of transportation from factory to site.

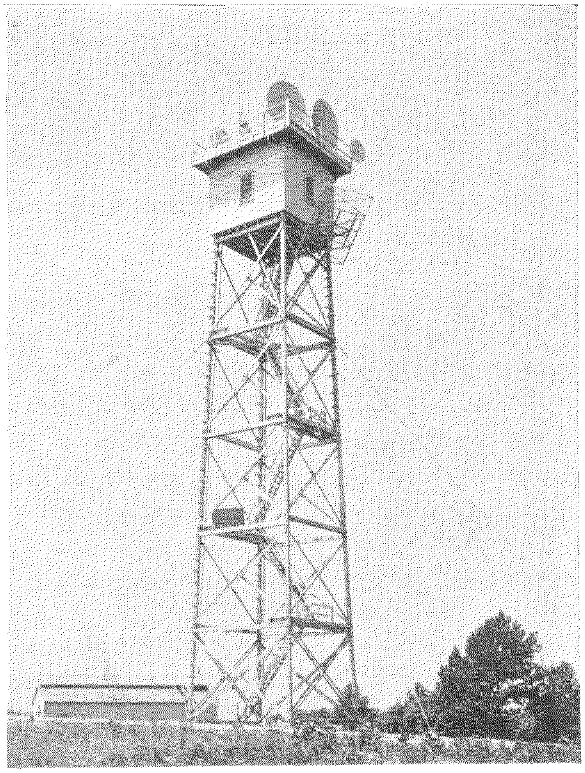


Fig. 16—100-foot wooden tower at Telegraph Hill, New Jersey.

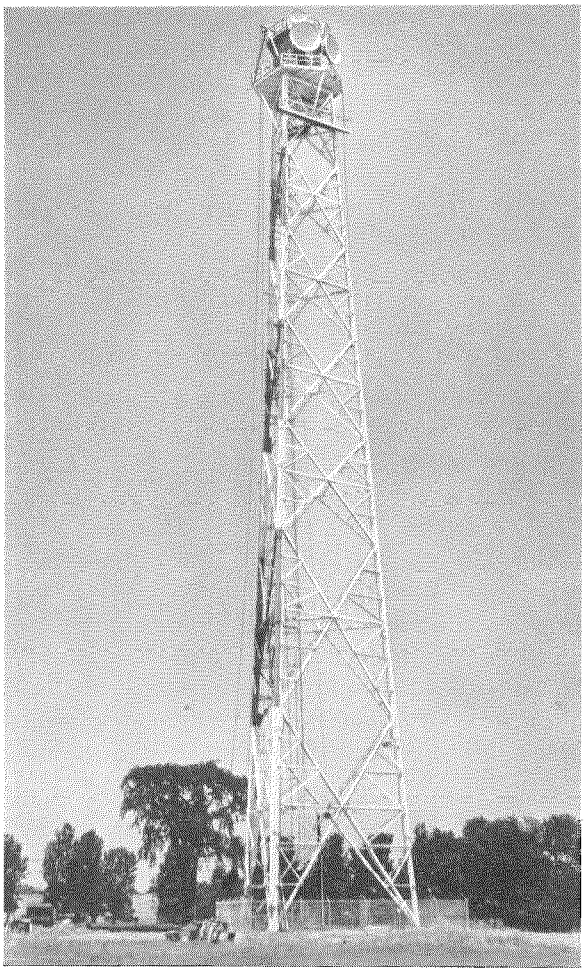


Fig. 17—200-foot wooden tower at Nutley, New Jersey.

Wood has some advantage in locations where it is a natural product. However, use of local material introduces an unknown factor of quality. Wood, of course, is also highly susceptible to damage by fire and lightning. It must be protected from insect and fungus attack, but does not require as frequent painting as steel.

It is also feasible to build a tower of concrete using local materials. This presupposes complete construction on the site, and hence limits the degree of quality control obtainable. Nevertheless, by proper safety factors, an entirely dependable structure can be built which is immune to fire, insect, and fungus. Figs. 16, 17, and 18 show representative towers that have been utilized in radio relays.

3.5 TRANSMISSION LINES

As mentioned previously, the method of supplying radio-frequency energy to the antenna depends to a large extent not only on the carrier frequency, but also on the location of the repeater apparatus with respect to the antenna. To simplify servicing, it is preferable that the complete repeater equipment be located at the base of the tower with the antenna at the top and some suitable connecting means such as a transmission line between them. In the experimental relay link described, the radio-frequency equipment was located at the top of the tower and a transmission line about 25 feet long was used. This short length, of course, greatly simplified the transmission-line problem and allowed the use of solid-dielectric coaxial cable.

For remote location of the radio-frequency equipment, rigid gas-filled lines are needed. An electrically operated compressor and dehydrator would be required to maintain proper pressure and keep the line dry. A more satisfactory solution is the use of a mirror arrangement on top of the tower with the antenna and radio-fre-

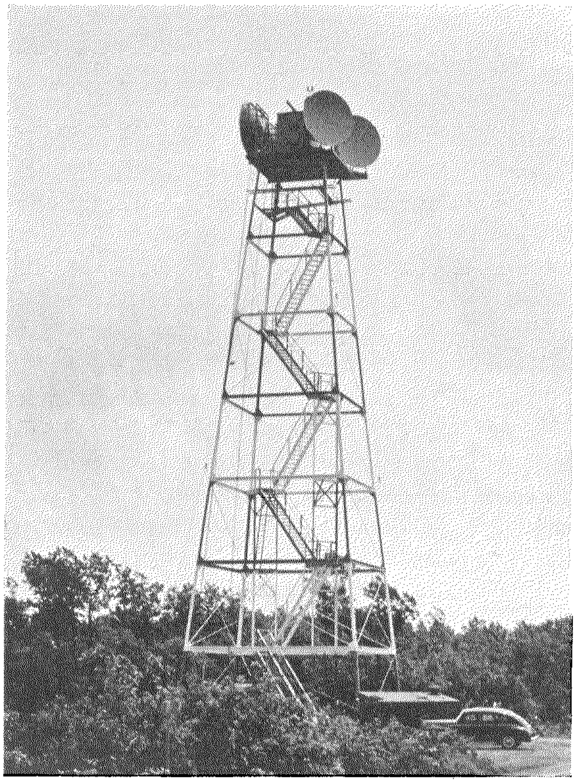


Fig. 18—100-foot steel tower at Laurel Hill, New Jersey.

quency equipment located at the base. Such an arrangement, however, is limited to the higher frequencies where a narrow beam can be projected with a reasonably sized antenna and mirror.

3.6 MISCELLANEOUS ITEMS

Items which comprise part of the relay plant but need be mentioned only briefly are the various housings to shelter and protect the radio-frequency equipments, antennas, and power units. In addition, there are the various control items such as relay units for switch-over of both radio-frequency and power-supply equipments and which also incorporate the necessary signaling apparatus to indicate at a remote monitoring location the condition of the equipment in the link. The characteristics of these items, of course, depend on the specific type of link as well as on the various operating features required.

4. Over-All Link Operation

4.1 GENERAL

To verify the performance of a radio link utilizing time-division multiplex with pulse-time modulation, experimental links incorporating the types of equipment previously described have been installed. Two such links have been set up; one having two repeaters with the terminals at a common location, arranged in a triangle, and a second representative of operation between two remote terminal points. As the closed-loop triangular link permits a simplified testing procedure, the majority of the measurements referred to have been performed on this link.

As indicated in Fig. 19, the terminals of the triangular link are at Broad Street, New York City, with one repeater at Telegraph Hill, New Jersey, and the second at Nutley, New Jersey. This arrangement includes an overwater link of 23 miles from New York to Telegraph Hill, an overland 30-mile path from Telegraph Hill to Nutley through a heavy industrialized area, and a comparatively short path of 10.6 miles to the southern tip of Manhattan island where the radio waves pass through gaps among the taller skyscrapers. The figure also shows the profiles of each path and the tower heights.

The repeater installations utilize the wooden towers illustrated in Figs. 16 and 17. Two 8-foot-diameter parabolas at each tower are connected by coaxial cable to the repeater equipment in adjacent shacks. Each station was operated from the local power main. The parabolas used at the Broad Street terminal were similar to those at the repeater points and the respective terminal radio-frequency transmitting and receiving equipment was located adjacently and connected by short runs of solid-dielectric coaxial cable. Telephone and pulse-time-modulated multiplex facilities were arranged at the common Broad Street terminal to provide two-way operation by assigning

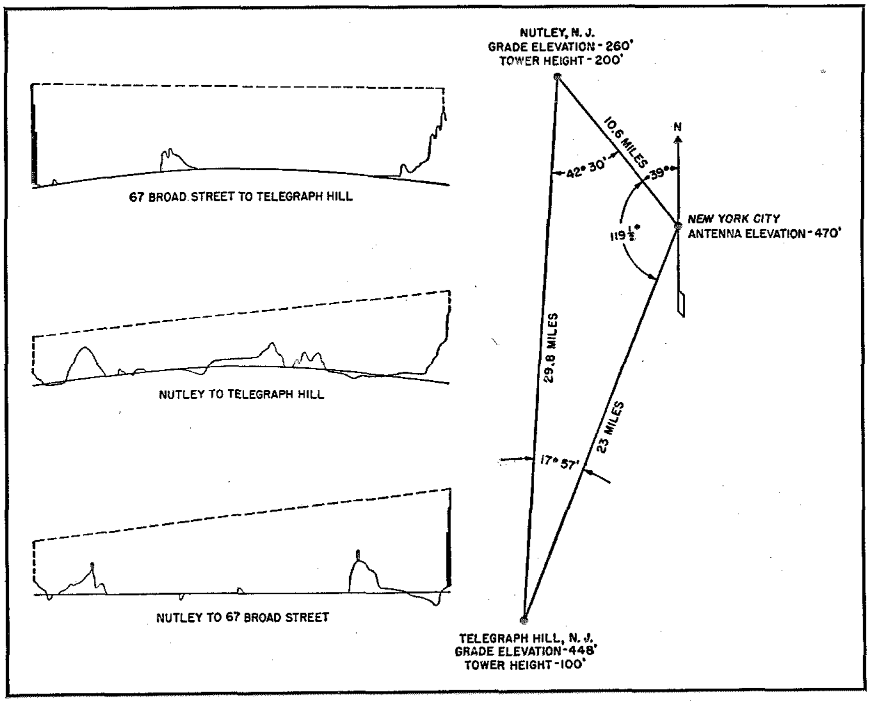


Fig. 19—Triangular relay system on which much experimental data were obtained. The terminal is at the southerly tip of Manhattan Island in New York City. Two repeaters are located at Nutley and Telegraph Hill, both in New Jersey. Overland and overwater links are thus included.

one half of the available channels for each of the calling and answering paths. In this manner, radio-frequency transmission was maintained in one direction while at the same time allowing simulation of two-way telephone operation. Two radio frequencies were employed; 1225 megacycles at the Broad Street transmitter, 1280 megacycles at the Telegraph Hill transmitter, and 1225 megacycles at the Nutley transmitter. They provided more than adequate frequency separation for the transmitter and receiver at each location.

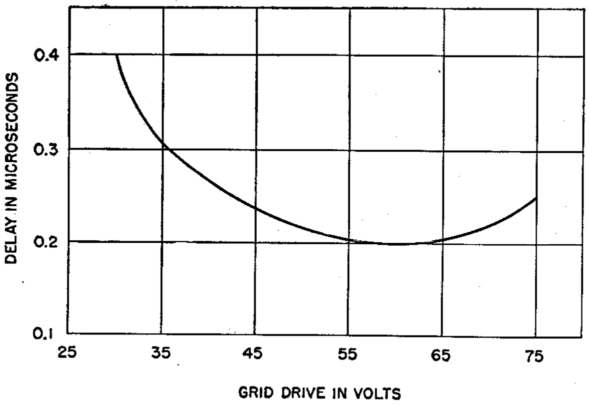


Fig. 20—Variation in starting time of the pulses as a function of the grid driving voltage on the oscillator. The pulse width was maintained constant.

4.2 Siting

Although the three antennas were nominally within line of sight of each other, conditions were such that alignment could not be achieved visually. By sighting prominent New York City skyscrapers from the Broad Street and Nutley locations, the true azimuth angles were determined and permitted approximate aiming of the antennas. Final readjustment was made after radio contact was established.

4.3 Signal to Noise

For the average power of 2.0 watts supplied by the radio-frequency equipment and for the signal-to-noise improvement ratio of 16 decibels provided by the pulse-time modulation method, the theoretical signal-to-noise ratio for the link was of the order of 53 decibels. Initial measurements indicated results to be short of the calculated value. It was found that inaccuracy in starting the transmitter oscillators reduced the signal-to-noise ratio unduly. The resistance noise in the oscillator was responsible for the variation in starting time. Several methods were found to minimize this noise.

One method utilized the trailing edge of the received pulse to start the transmitter as the noise on this edge is less than that on the leading edge. A second method involved the injection of radio-frequency power into the oscillator during pulses to provide a "catalyzation" effect or, alternatively, a similar effect was obtained by maintaining a small constant plate voltage between pulses. Fig. 20 illustrates the variation in starting time as a function of the oscillator operating conditions. The simplest method, and the one utilized in the link equipment, involved increasing the oscillator excitation by adjustment of the grid and cathode coupling. Under this condition, the transmitter noise was reduced to more than 65 decibels below the signal and thereby introducing a negligible amount of noise into the link. With this reduction in transmitter noise, it was found possible to approach closely the theoretical signal-to-noise ratio for average propagation conditions.

4.4 Multipath Reflection

The effect of multipath reflection on the various methods of transmission is similar. With frequency modulation, for example, distortion is experienced and may give rise to cross-talk effects if multichannel signals are transmitted. A similar effect can be obtained with pulse transmission where time-division multiplex is utilized. If the secondary wave is delayed from the primary wave by a time equal to the timing between adjacent channels, the information corresponding to the two channels arrives simultaneously and cross talk results. This effect is, of course, reduced considerably by the directivity of the transmitting and receiving antennas and also if the antenna locations are suitably chosen.

Although some difficulty with this effect was anticipated, the cross talk experienced in the link was found to be negligible even under conditions where antenna locations were unfavorable and where large reflecting objects such as buildings and hills were located adjacent to the

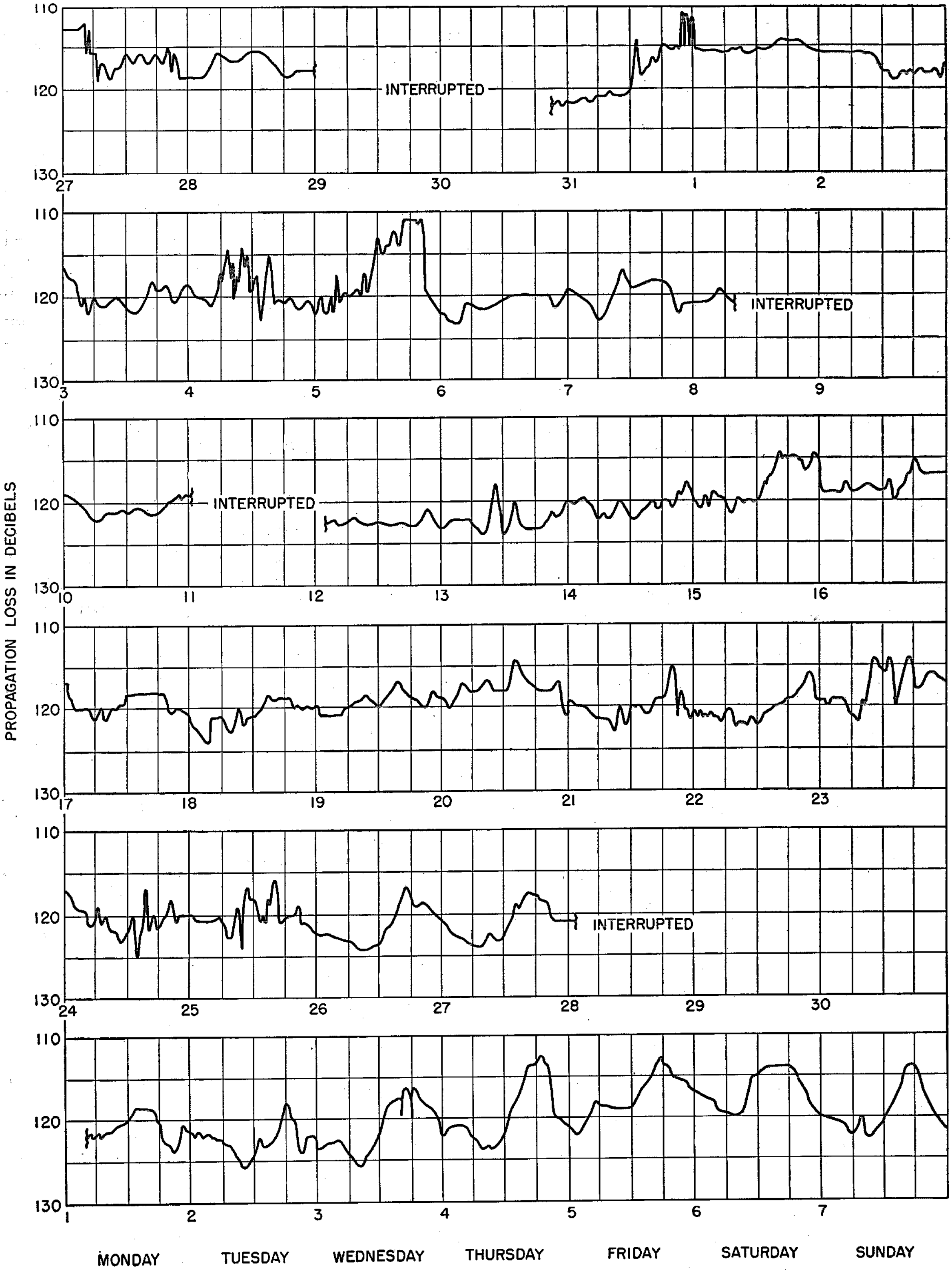


Fig. 21—Signal variations over the Nutley—Broad Street path during representative periods of a 42-day test starting on Monday, May 27 and terminating on Sunday, July 7, 1946. Propagation loss is the relation between the transmitted and received powers.

repeater sites. The absence of cross talk can be ascribed to the directivity of the antennas and the interference-reducing properties of the transmission method. In this latter case, the use of limiters and similar circuits, together with the noise-improvement ratio of the system, minimized the effects of the resulting carry-over between pulses.

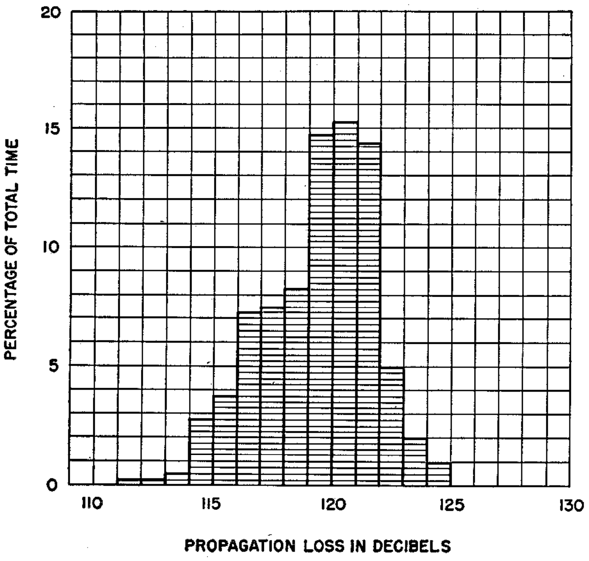


Fig. 22—Statistical presentation of the data given in Fig. 21.

4.5 PROPAGATION VARIATIONS

To check propagation variations, continuous 24-hour-a-day records were made at the receiver outputs along the link. Fig. 21 illustrates the observed signal variations over the Nutley–Broad Street path for representative periods during a 42-day interval. The statistical distribution of attenuation for the data given is plotted in Fig. 22. These charts show a variation of ±6 decibels from the theoretical attenuation of 119 decibels for this path.

It should be noted that the path is essentially an all-land route. Tests on the overwater link, Telegraph Hill to New York, have shown more severe fades of the order of 20 decibels. These fades were particularly noticeable during hot damp summer days when no wind was present. This would tend to indicate a change in the relative phase between the free-space and the reflected vectors resulting from a change of velocity of the reflected wave through the water-laden air near the water surface. For the normal fades, no dropouts were experienced although momentary loss of signal was observed during severe fading conditions. Although diversity reception was not utilized, it would be expected that a minimization of the rapid-fading dropouts would be obtained by this means. Negligible effects were observed from snow, ice, and rain, as well as seasonal variations except for the conditions previously indicated.

4.6 GENERAL OBSERVATIONS

Although the equipment was not specifically designed for continuous operation, it was possible to obtain essentially unattended operation at the repeater points after initial adjustments were accomplished. It was thus possible to operate the radio relay in a manner similar to that of a normal wire line. Interruption of service caused by propagation anomalies was rather a rare occurrence and reliable telephone service on an experimental basis was possible.

The equipment described is in operation and additional over-all tests are being conducted at present. To determine the full applicability of such relays to commercial practice, a large amount of testing will be required. It is safe to anticipate, however, that a time-division-multiplex relay similar to the type described will offer attractive possibilities as a commercial multichannel radio relay system.

5. Acknowledgment

Acknowledgment is made to Mr. E. Labin, Mr. H. G. Miller, and various engineers and technicians of Federal Telecommunication Laboratories who contributed to the development and design of the equipment described.

Paris-Montmorency 3000-Megacycle Frequency-Modulation Radio Link*

By A. G. CLAVIER and G. PHELIZON

Laboratoire Central de Télécommunications, Paris, France

INAUGURATION by the Minister of Posts, Telegraphs, and Telephones of a centimeter-wave link between Paris and Montmorency occurred on April 19, 1946. This is the first time that a multiplex radio link has been inserted in the telephone network of the Paris region. The new service permits simultaneous transmission of twelve conversations in each direction, with a quality of transmission comparing favorably with the best type of telephone cables.

The quality demanded of international telephone circuits corresponds to a ratio of signal to intelligible cross talk of at least 65 decibels in each audio-frequency channel. This is the main technical difficulty encountered in inserting a multiplex radio section in a high-quality telephone line. The cross talk originates with nonlinearity in the modulation stages at the transmitting end, and in the demodulation stages at the receiving end. Special circuits were designed to achieve the desired magnitude of signal-to-cross-talk ratio.

There are many advantages in certain cases in substituting a radio link for a section of telephone cable. The obvious instance is where considerable difficulties are encountered in installing and maintaining submarine telephone cables. It is natural that such insular countries as Great Britain and Japan were the first to apply very high radio frequencies to multichannel telephone communication, to bridge bodies of water 10 to 80 kilometers wide. As a forerunner, the communication established in 1933 between the airports of Lympne, England, and Saint Inglevert, France,[1] is worthy of mention. A wavelength of 17 centimeters (1800 megacycles per second) was used, and the radiation was focussed in a sharp beam by parabolic reflectors; one telephone or teleprinter circuit was operated across the English Channel until the beginning of the second world war.

A commercial link was put in operation in 1937 between Scotland and Ireland, with 9 telephone channels on a wavelength of 4 meters (75 megacycles). The distance is 65 kilometers with line-of-sight transmission. This is an amplitude-modulation system, and the signal-to-noise ratio is approximately 45 decibels in each channel. Since that time, other links have been installed utilizing amplitude modulation; the cross-talk performance has been improved by using inverse feedback.

Centimeter waves are inherently suitable for multiplex radio systems, as had been foreseen when the Lympne–Saint Inglevert link was designed, and still earlier in the preliminary Calais-Dover[2] experiments of 1930. A high concentration of radiated energy in a desired direction is readily achieved with microwaves, the antennas for which are of relatively small dimensions. Also, wider frequency bands are easily obtained and handled.

This latter fact makes the use of frequency modulation attractive. The radio transmitter may be frequency modulated with the signal from a standard carrier cable system utilizing frequency-division multiplexing. Frequency modulation brings a number of well-known advantages, such as an improvement of signal-to-noise ratio proportional (in root-mean-square value) to the modulation index; independence from nonlinearity of vacuum-tube characteristics; and stability over a wide range of received signal strength through the use of amplitude-limiting devices.

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* Originally published in French as "Le Câble Hertzien," *L'Onde Electrique*, v. 26, pp. 331–334; August-September, 1946.

[1] A. G. Clavier and L. C. Gallant, "Anglo-French Micro-Ray Link Between Lympne and St. Inglevert," *Electrical Communication*, v. 12, pp. 222–228; January, 1934.

[2] "Micro-ray Radio," *Electrical Communication*, v. 10, pp. 20–21; July, 1931.

To achieve the required cross-talk level, however, it has been found necessary to add carefully designed feedback circuits both at the transmitter and receiver. Inverse feedback is already well known in carrier-cable amplifiers and in some amplitude-modulation radio circuits. When applied to frequency-modulation equipment, it takes a particular aspect, inasmuch as it results in a compression of frequency deviation in the intermediate-frequency circuits of the receiver.

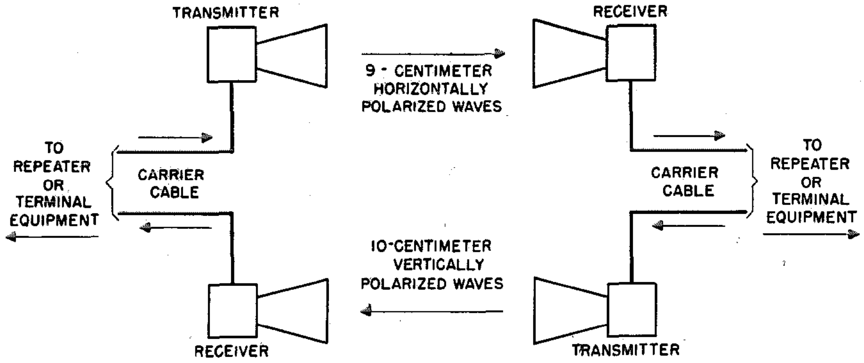

TRANSMITTER
RECEIVER
9 - CENTIMETER HORIZONTALLY POLARIZED WAVES
TO REPEATER OR TERMINAL EQUIPMENT
CARRIER CABLE
CARRIER CABLE
TO REPEATER OR TERMINAL EQUIPMENT
10-CENTIMETER VERTICALLY POLARIZED WAVES
RECEIVER
TRANSMITTER

Fig. 1—General arrangement of the Paris-Montmorency radio link.

The improvement attained through inverse feedback is known to be limited by the degree of fidelity achieved in the feedback loop. In frequency modulation, the feedback loop includes a local oscillator, the frequency of which must vary in a very linear way with the applied voltage. This constitutes the limiting "linear reference" of the system. Fortunately, the positive-grid oscillator used in the Calais-Dover experiments has a very linear characteristic over a frequency range of some 50 megacycles. Used for a frequency deviation of only ±0.5 megacycle, it is sufficiently distortion-free to permit the severe international cross-talk standard to be achieved.

The design of the Paris-Montmorency equipment was preceded by experiments carried on during the last months of 1941 near Toulon, on the Mediterranean coast. The purpose of these experiments was to test the properties of horn-type directional aerials and to study the propagation of 3000-megacycle waves. The transmitter was located north of Toulon at an altitude of 700 meters. A receiver on a ship made it possible to test the propagation of high-frequency signals to the horizon and beyond. A number of experiments revealed that beyond the horizon, two modes of propagation were encountered; the normal one with a fast quasi-exponential decrease of field strength and, under certain atmospheric conditions (mostly hot and calm weather), an anomalous one in which the field strength passed through several minimum and maximum values out to a distance of more than 80 kilometers beyond the horizon of the transmitter.

Frequency-modulated waves were used and the limiter held the audio-frequency signal constant over a wide variation of high-frequency field strength.

Consideration was then given to the design of a radio equipment to be inserted as a section of a long-distance telephone line; this led to the Paris—Montmorency installation.

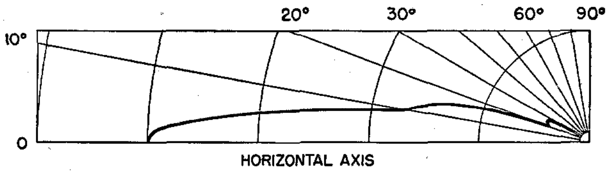


Fig. 2—Polar diagram of vertically polarized radiation from the horn in the vertical plane. The beam angle is about 11 degrees at points 3 decibels below maximum relative voltage.

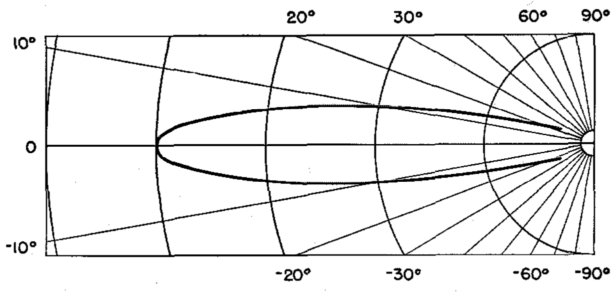


Fig. 3—Vertically polarized radiation from the horn in the horizontal plane. The beam angle is 14 degrees at the 3-decibel points of relative voltage.

1. General Characteristics

The signal to be transmitted comes from the carrier terminal equipment and consists of 12 audio-frequency channels transposed in frequency within a band ranging from 12 to 60 kilocycles. A 4-kilocycle band is allocated to each channel.

The general arrangement of the system is shown in Fig. 1. It can be said to be of the "4-wire" type, a separate carrier frequency being used for each direction of transmission. The Paris-Montmorency direction operates on 3333 megacycles (9 centimeters), and the opposite direction utilizes a 3000-megacycle wave (10 centimeters). As an added precaution, the two planes of polarization are at right angles; horizontal from Paris to Montmorency, and vertical for the return direction.

The composite signal from the 12-channel telephone cable is applied directly to modulate the transmitter. Conversely, at the receiving end, the signal from the radio receiver is introduced directly into the cable.

The directional radiators are of the horn type. Because of the rather high transmitter output (30 watts), the aperture area need only be 0.5 square meter for satisfactory reception. This results in a gain of about 22 decibels for the 3000-megacycle wave, as compared with a half-wave antenna in free space. The gain is somewhat larger for the 9-centimeter wave. The horn is preferred to a parabolic reflector because of its sturdiness and the ease with which it can be constructed. The problem of coupling the transmitter output to the radiator is also simpler and more efficient, as the transmission line is at the rear of the horn, instead of being in front as in the case of a reflector.

Figs. 2 and 3 show the shape of the radiation patterns as experimentally determined. Different flare angles have been adopted for the two sides of the horn; for instance 25 degrees in the vertical plane, and 30 degrees in the horizontal plane, for vertical polarization. As the diagrams show, lateral lobes have thus been avoided. The sharpness of the beam is such that the 3-decibel attenuation points occur at ±6 degrees in the vertical, and ±7 degrees in the horizontal planes.

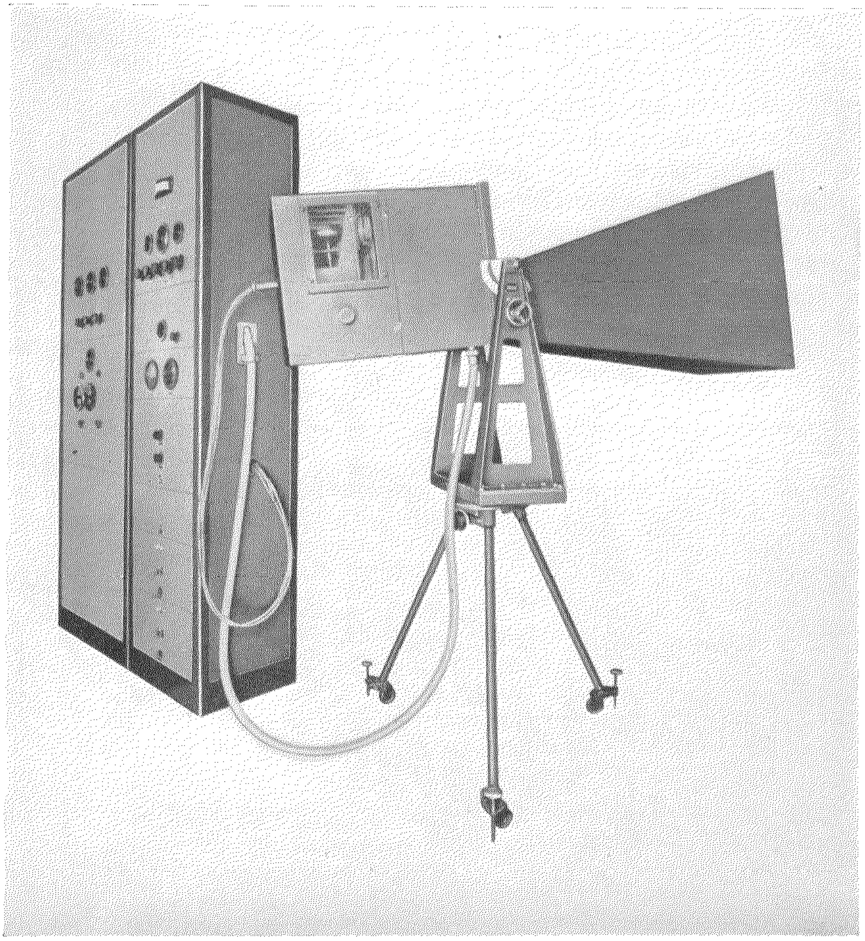


Fig. 4—Transmitting equipment. The microwave transmitting tube and radiating horn are mounted on the tripod for convenience in aiming.

The small end of the straight-sided horn terminates in a rectangular wave guide which is coupled to the transmission line to the transmitter or receiver. The waves in the guide are of the $H_{0,1}$ type, either vertically or horizontally polarized as required. Each horn is mounted on a tripod fitted with graduated sectors for the exact adjustment of elevation and azimuth of the beam axis.

2. *Transmitting Equipment*

Fig. 4 shows a general view of the transmitting equipment. A velocity-modulated oscillator delivers about 30 watts to the transmission line. This tube, developed in 1941, is placed in a housing mounted on one side of the smaller end of the horn, as shown in Fig. 5. The output of the oscillator is transmitted to the horn through a coaxial line. An impedance-matching arrangement is provided and may be seen in Figs. 5 and 6.

Frequency modulation of the oscillator is obtained by variation of the voltage applied between cathode and cavities of the tube. It is particularly important to maintain this 5000-volt supply constant, as fluctuations are an obvious source of noise. Therefore, a fraction of the ouput voltage is compared to the fixed voltage of a dry battery, and the difference is applied to the input of a direct-current amplifier. The output circuit of the amplifer is in series with the power supply circuit and thus stabilizes the output voltage. The residual ripple is less than 1 part in $3 \cdot 10^4$ of the applied voltage. The stabilization ratio, defined as the ratio of the percentage variation of the mains voltage to the corresponding percentage variation of the regulated voltage, is approximately 1000. With this arrangement, the frequency

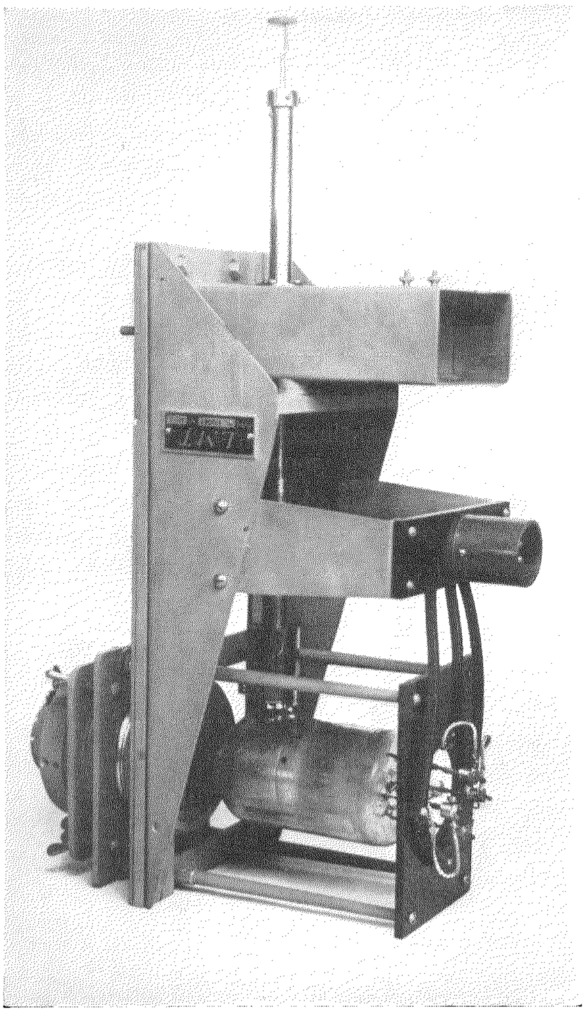


Fig. 6—Oscillator and transmission lines to the horn with their adjustment stubs.

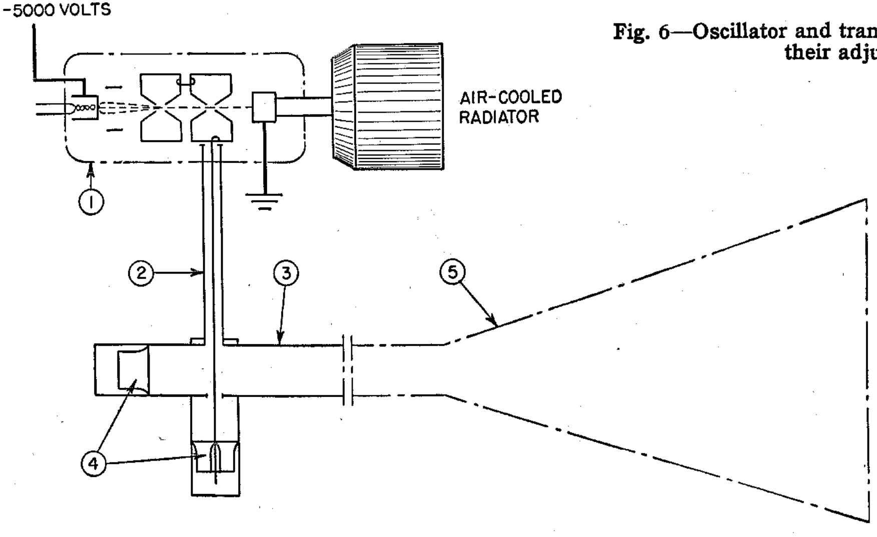


Fig. 5—Arrangement of oscillator and radiator. 1 is the velocity-modulated oscillator, 2 is the transmission line which connects the oscillator to the rear section of the horn, 3 is a section of wave guide to which the horn is coupled, 4 are the tuning stubs, and 5 is the horn radiator.

stability of the transmitter is of the order of 1 part in 10^5.

The maximum frequency of the signal being 60 kilocycles, a modulation index of 6 has been adopted, leading to an improvement of approximately 20 decibels in signal-to-noise ratio compared with amplitude modulation. The maximum frequency deviation is thus ±360 kilocycles, and the voltage-frequency characteristic of the tube is sufficiently linear for a range of ±500 kilocycles.

The modulating signal is received at the input level normally applied to a repeater for a carrier cable. A modulation amplifier must be inserted, and distortion is kept very low by use of inverse feedback. The signal-to-distortion ratio of the transmitter was found to be of the order of 50 decibels. This is not sufficient, and to obtain the specified 65-decibel ratio, frequency-modulation feedback circuits were developed.

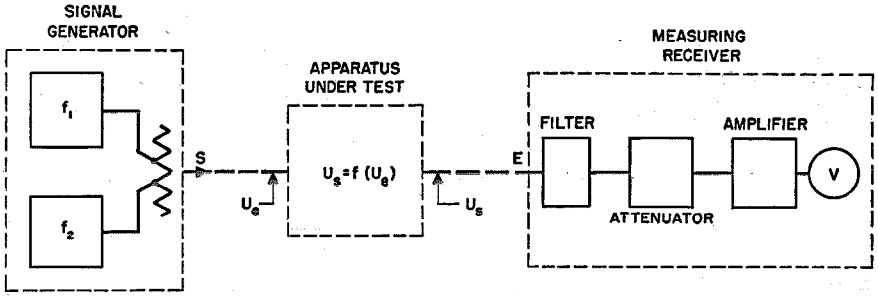


Fig. 7—Schematic arrangement of equipment for measuring cross talk using two input signals.

2.1 MEASUREMENT OF DISTORTION

It may be of interest to describe briefly the method and instrument used to measure over-all and component distortions, so as to determine the improvement actually contributed by the frequency-deviation-compression arrangements.

The block diagram of the measuring apparatus is shown in Fig. 7. Two sine-wave voltages of frequencies f_1 and f_2, which are in the modulating range of 12 to 60 kilocycles, are produced by a signal generator having the very low distortion level of −75 decibels. These signals are applied with equal amplitudes to the input of the equipment to be checked. At the output of the equipment, there will then appear, in addition to the two input signals, parasitic terms originating in the equipment as a result of nonlinearity.

The nonlinear distortion can be evaluated by comparing the amplitudes of the fundamental frequencies with the level of the so-called 2nd-degree terms, such as $2f_1$, $2f_2$, f_1-f_2, etc., and of the 3rd-degree terms, such as, $3f_1$, $3f_2$, $2f_1$-f_2, etc. This is achieved by a special measuring receiver tuned to a fixed frequency of 20 kilocycles, and provided with means of measuring 10 microvolts accurately. Accurate measurements of distortion terms of the order of −65 decibels are thus possible. The adjustment of separate elements and of the entire installation made in this manner have been confirmed by standard measurements in the low-frequency output channels.

2.2 FREQUENCY-MODULATION FEEDBACK CIRCUITS

Just as in the case of amplitude modulation, the distortion of a direct path of amplification can be reduced by the addition of an inverse path. This corrects the defects of the direct path provided that distortion does not occur in the return path. To achieve the desired result, the return path itself had to be fitted with an inverse-feedback loop, and thus the total arrangement includes both a local and an over-all feedback loop, as shown in Fig. 8.

Frequency-modulation feedback circuits can, of course, operate only if the loop circuits are sufficiently stable. The general principles are the same as those given in the fundamental works of Black, Nyquist, and Bode dealing with amplitude-modulation feedback circuits. The stability of the amplifier is similarly evaluated by means of the "Nyquist criterion." The modulus of the amplification factor around the entire loop must fall below 1 at frequencies for which the phase shift reaches 0 or 2π, the phase shift being substantially equal to π in the useful band where inverse feedback is utilized. To obtain a stable inverse-feedback amplifier, it is necessary to control the gain characteristic around the complete loop from the highest useful frequency to

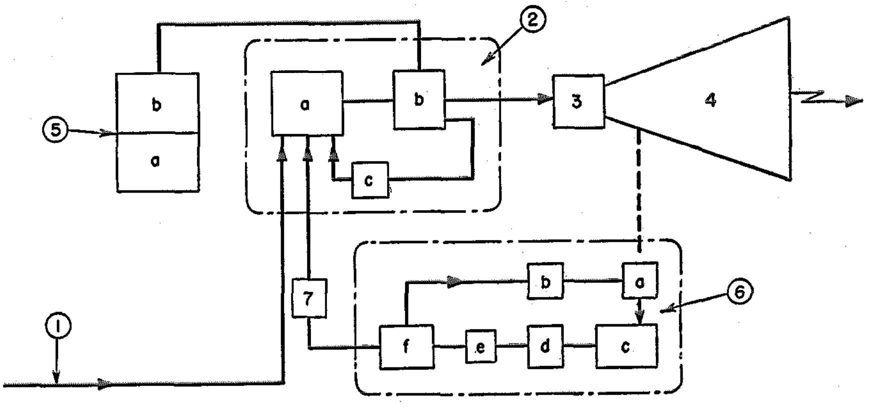


Fig. 8—Schematic arrangement of the feedback circuits in the transmitter. 1, carrier cable; 2, *a* and *b* are modulation amplifiers and *c* is a phase corrector; 3, transmitting tube; 4, radiating horn; 5, stabilized high-voltage supply; 6, auxiliary feedback receiver consisting of *a* mixer, *b* positive-grid local oscillator, *c* 9±1-megacycle intermediate-frequency amplifier, *d* limiter-discriminator, *e* phase corrector, and *f* direct-current amplifier; and 7, phase corrector for the over-all feedback loop.

the asymptotic part of the curve, which is determined at the higher frequencies by the unavoidable parasitic capacitances in the tube and circuits employed. There is thus an optimum cutoff characteristic which should be followed as closely as is practicable up to a frequency which increases with the inverse-feedback ratio and with the number of amplification stages within the loop. Consequently, amplifiers must be designed with a band width considerably larger than the useful band, and passive phase-correcting networks generally have to be added.

The best way to determine the safety margins that separate the actual characteristic from the conditions that would cause oscillation is to develop the Nyquist diagram; the gain around the open loop is plotted as a complex vector in a system of polar coordinates. This, however, means measuring not only the absolute value of gain, but also the phase shift around the entire loop, which requires the use of a phase meter capable of operating at relatively high frequencies. It has been found more convenient to avoid direct measurement of phase shift and operate as follows, provided the looped amplifier has already proved to be stable.

In Fig. 9, let a gain-measuring apparatus be inserted in the inverse-feedback loop as shown, and let A, B, and C be the complex values of amplification for the forward path, the return path, and the over-all amplifier, respectively. Those three complex values are related by

$$C=\frac{A}{1-AB};$$

that is to say,

$$AB-\frac{A}{C}=1.$$

In the latter expression, AB represents the complex gain around the open loop, and A/C represents the ratio of the gain of the amplifier before and after the introduction of inverse feedback. Let the moduli of AB and A/C be measured; it is then possible to trace the triangle representing the above equation and to deduce the value of the vector AB. The Nyquist diagram can thus be plotted as in Fig. 9, and the degree of stability of the amplifier can be evaluated. This determination is accurate only if AB is approximately 1, the region in which a precise knowledge of the diagram is important.

To obtain an improvement in signal-to-noise ratio of 20 decibels over amplitude modulation for a frequency band of 60 kilocycles with a large number of stages of amplification, the shape of the gain characteristic had to be investigated up to 1 megacycle. The intermediate-frequency amplifier of the auxiliary frequency-modulation receiver has a bandwidth of 2 megacycles; the modulation amplifier has constant gain up to 1 megacycle. Suitably connected phase-correcting circuits are introduced to obtain optimum cutoff characteristics.

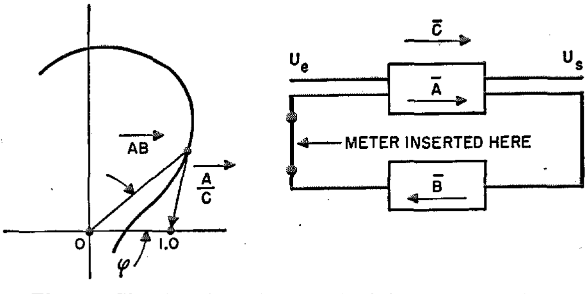


Fig. 9—Simplified method of obtaining data for plotting Nyquist diagrams. Only measurements of gain are required.

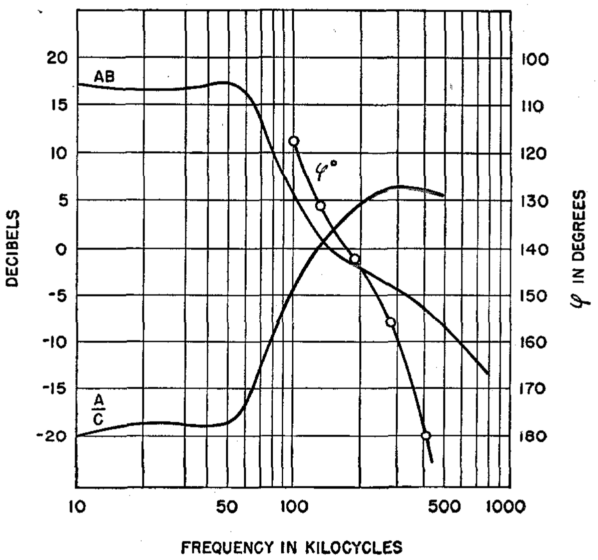


Fig. 10—Curves for determining the general frequency-compression characteristics of the transmitter.

As already mentioned and shown in Fig. 8, there are two main feedback loops in the transmitter, not counting the additional amplitude-modulation-feedback loop in the modulation amplifier. One is a so-called "local loop" which affects the auxiliary frequency-modulation receiver only; the second one is the over-all loop, which concerns the whole of the transmitter from radio-frequency output back to the low-frequency input at the modulation amplifier.

The auxiliary receiver is a superheterodyne, with a local oscillator of the positive-grid type having a very linear relation between frequency deviation and anode voltage. The mixer is a special diode with short electron transit time between cathode and anode. The frequency converter is followed by an intermediate-frequency amplifier, limiter, and frequency discriminator. The modulating signal is thus locally reproduced and applied to the anode of the receiver local oscillator in a direction suitable to cause a compression of the frequency deviation at the receiver output. As a result, there is a considerable decrease in the distortion which usually attends demodulation. In addition, the local-feedback loop being arranged to transmit the direct-current component, the center value of the local-oscillator frequency follows any drift of the transmitter, and the intermediate-frequency signal remains within the pass band of the receiver.

The auxiliary receiver described has a high degree of fidelity and can be incorporated as the return path in the general frequency-deviation-compression circuit of the transmitter. The output of this receiver is applied to the input of the modulation amplifier in phase-opposition to the main signal, so that inverse feedback is obtained. This is effected through a phase-correcting circuit which adjusts the gain around the open loop to approximately 17 decibels in the 12- to 60-kilocycle band, and produces a cutoff characteristic above 60 kilocycles such that the stability of the whole circuit is assured with a sufficient margin of safety. The curves shown in Figs. 10 and 11 illustrate the results actually obtained.

The improvement in quality of transmission resulting from frequency-modulation feedback has been confirmed by means of the cross-modulation meter described previously. Accordingly, the signal-to-distortion ratio, which is of the order of 50 decibels without frequency compression, has been found to exceed 65 decibels for full modulation of the transmitter when the feedback arrangement is utilized.

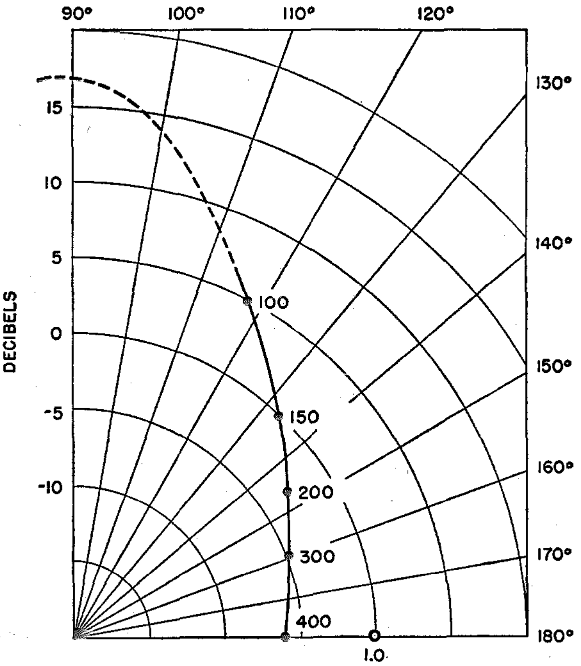


Fig. 11—Nyquist diagram of the frequency-modulation feedback circuit of the transmitter.

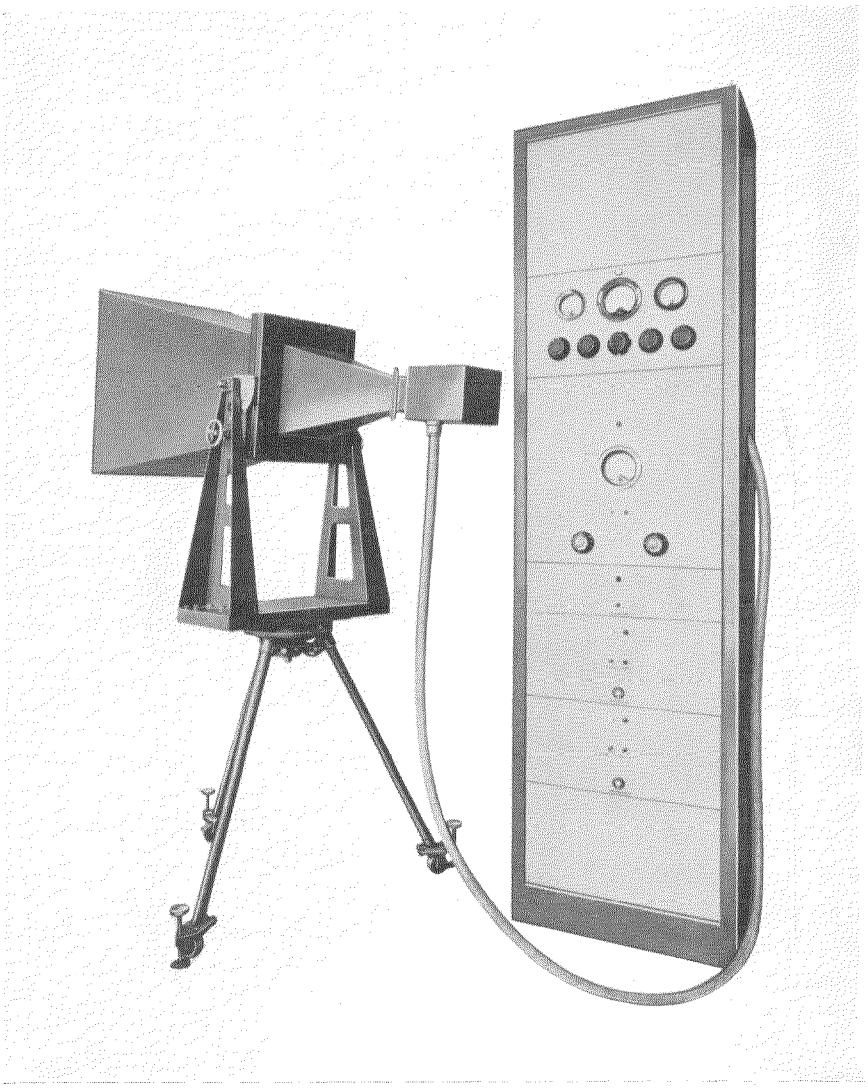


Fig. 12—9-centimeter receiver of the Paris-Montmorency radio link.

3. *Receiving Equipment*

At the receiving end shown in Fig. 12, the high-frequency signal is picked up by a horn identical to that of the transmitter. From the rear wave-guide portion of the horn, the signal is sent through a coaxial cable to the mixer. Use is made of a short-transit-time diode, and a positive-grid local oscillator. The intermediate-frequency signal is amplified at 15 megacycles, and then converted to 9 megacycles. The total intermediate-frequency gain is 120 decibels over a 1-megacycle band. A limiter-discriminator detects the frequency-modulation signal and the low-frequency output is sent to the carrier cable terminals.

Fig. 13 is the block diagram of the receiver. Note that it is provided with a feedback loop for frequency-deviation compression. To achieve this, the output signal from the discriminator is applied to the anode of the positive-grid local oscillator in such a direction as to reduce the frequency deviation in the receiver intermediate-frequency circuits. This loop includes a phase corrector; stability is adjusted and checked by the method already described for the transmitting equipment.

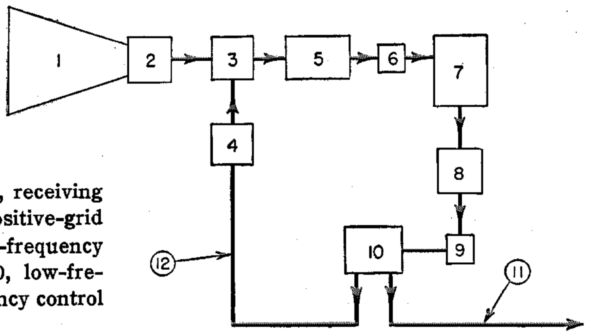


Fig. 13—Schematic arrangement of the receiver. 1 and 2, receiving horn and matching circuits; 3, balanced diode mixer; 4, positive-grid local oscillator; 5, 6, and 7, 15- and 9-megacycle intermediate-frequency amplifiers; 8, limiter-discriminator; 9, phase corrector; 10, low-frequency amplifier; 11, carrier cable; and 12, automatic-frequency control and feedback loop.

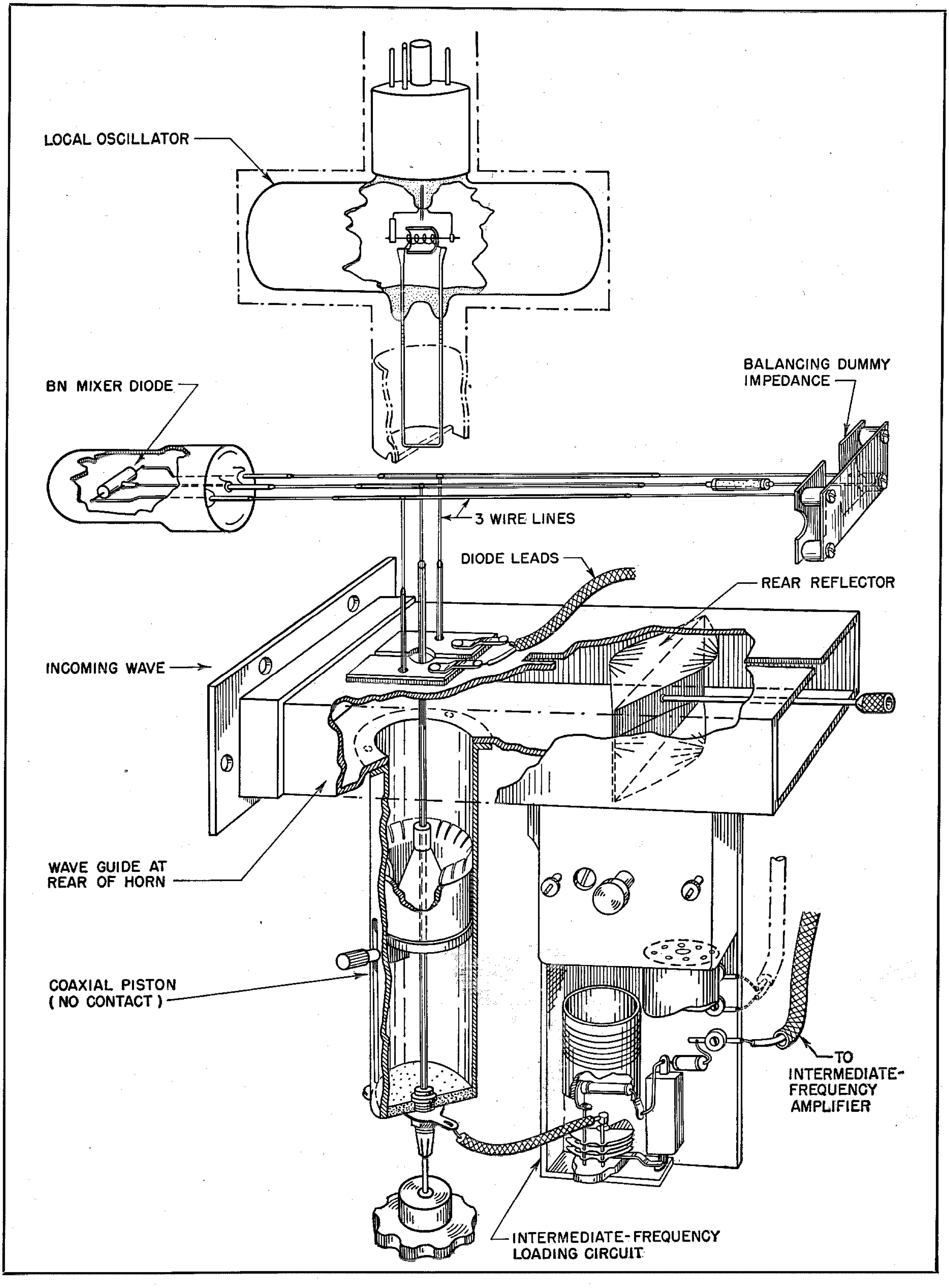

LOCAL OSCILLATOR
BN MIXER DIODE
BALANCING DUMMY IMPEDANCE
3 WIRE LINES
DIODE LEADS
REAR REFLECTOR
INCOMING WAVE
WAVE GUIDE AT REAR OF HORN
COAXIAL PISTON (NO CONTACT)
TO INTERMEDIATE-FREQUENCY AMPLIFIER
INTERMEDIATE-FREQUENCY LOADING CIRCUIT

Fig. 14—Balanced mixer used in receiver. This entire equipment is mounted in the box at the rear of the horn as shown in Fig. 12.

The frequency deviation of the positive-grid oscillator with variation of voltage must be very linear in the range of frequency deviations utilized, as this is the limiting factor in freedom from cross talk. It has been found that when such an oscillator is coupled directly to the receiver horn, reflections back into the horn from exterior obstacles cause irregularities in the operation of the tube. It is, therefore, necessary to do away with any possible radiation from the local oscillator. To achieve this, the tube is mounted in a shielded housing and the mixer circuit is of the balanced type, balance being obtained through the use of a dummy impedance duplicating that of the mixer diode as shown in Fig. 14. It has thus been possible to decrease the radiation of the local oscillator by approximately 30 decibels, and under these conditions the linearity of the tube characteristic remains practically unimpaired.

It has been shown above that the intermediate-frequency amplifiers should be carefully adjusted to have bandwidths of one megacycle for the receiver, and two megacycles for the transmitter auxiliary receiver. Furthermore, the discriminators must have operating bandwidths at least as large as the bandwidths of the associated intermediate-frequency amplifiers, and their demodulation characteristics must be as linear as possible. Adjustment of all of these circuits is difficult and lengthy if each point has to be plotted; the process is greatly simplified by projecting the whole response curve on an oscilloscope during adjustment. To do this, a signal generator must be obtained which can produce a frequency-modulated signal having constant amplitude over a range of several megacycles. One convenient way of doing this is to use the beat frequency between two positive-grid oscillators, one being stabilized by a high-Q cavity, and the other being modulated with a low-frequency generator by variation of its anode voltage. Fig. 15 shows oscillograms obtained by this method from the auxiliary receiver of the transmitting equipment.

4. *Conclusion*

The entire equipment of the Montmorency terminal is shown in Fig. 16. The installation has been in traffic operation since January, 1946, and has given very satisfactory results. This is a first step in the insertion of radio links in high-quality multiplex telephone networks. The number of channels will, of course, have to be increased considerably to compete with cables. The transmission range will also have to be extended; the Paris-Montmorency link is only 15 miles long, while the same equipment could be used for twice that distance provided the terminal stations were located in sight of each other. Further extension may be provided by the installation of relay stations. There is no doubt that the recent advances in centimeter-wave technique will soon be applied to covering longer distances and supplying a greater number of simultaneous channels. The installation inaugurated at Paris will nevertheless remain a pioneering step in that direction.

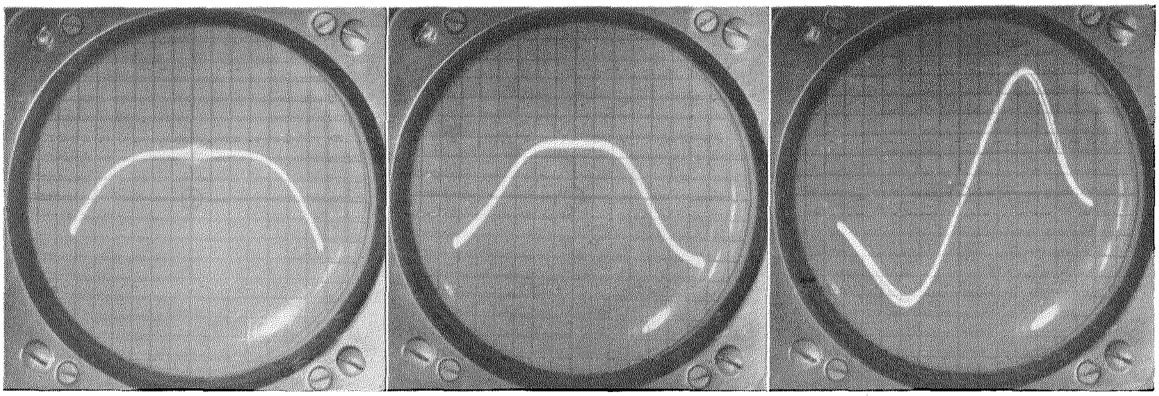


Fig. 15—Three oscillograms obtained from the auxiliary receiver of the transmitting equipment. The first two curves are the response-frequency characteristics of the intermediate-frequency amplifiers. The bandwidths are 2 megacycles. The discriminator response-frequency characteristic is displayed in the oscillogram at the right.

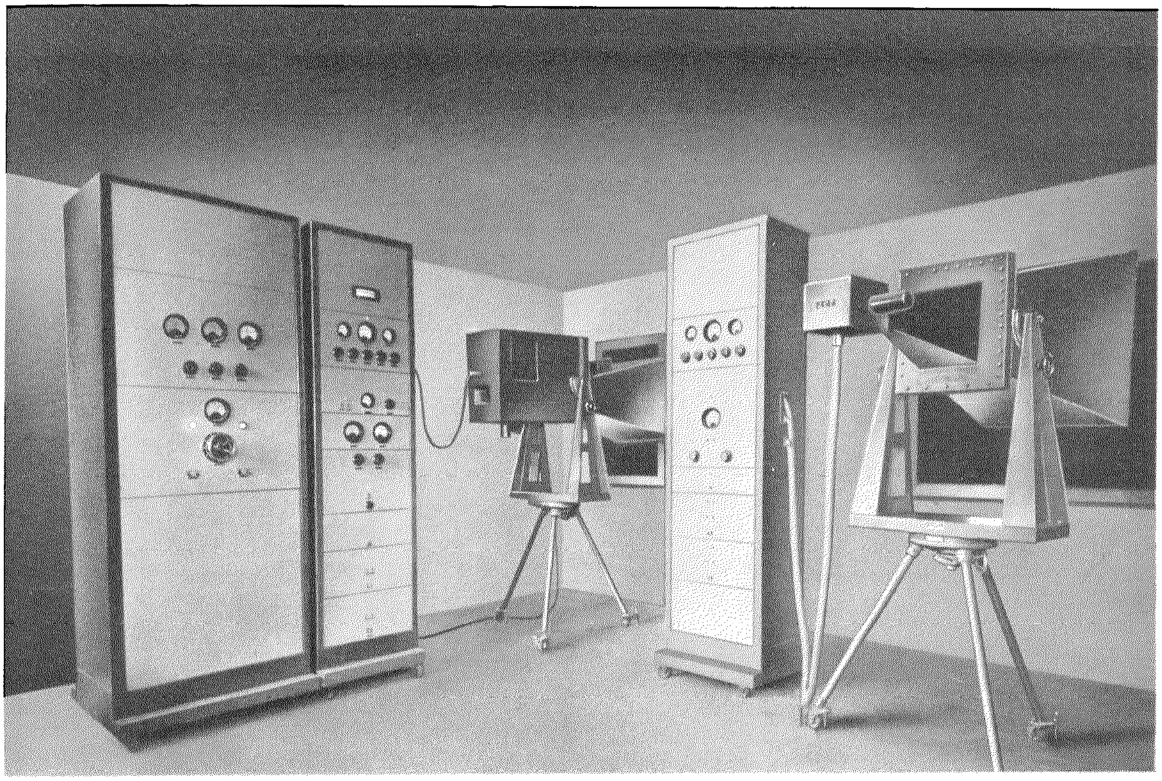


Fig. 16—Montmorency terminal. The 9-centimeter receiving horn is at the right, and the 10-centimeter radiating unit is in the center. The multiplexing equipment and power supplies are contained in the three cabinets.

5. *Acknowledgment*

The successful conclusion of this project was made possible by the cooperation of the French governmental technical services and the research laboratories of private industry. It may be interesting to note that this work was conducted in France during the most difficult period of enemy occupation. Started in 1941, it was actively helped by the French Communications Coordination Committee, of which Admiral Bourrague was President, and to which the late Colonel Labat gave his valuable cooperation. It was subsequently sponsored by the Administration of Posts, Telegraphs, and Telephones, under Mr. Lange, Director, and Mr. Marzin, Assistant Inspector General and Director of the Research Center.

A group of engineers and technicians of Laboratoire Central de Télécommunications contributed to the success of the project. The authors of the paper express their thanks to all their colleagues, and particularly to Mr. Altovsky.

Mobile Frequency-Modulation 30–44-Megacycle Equipment

By R. B. HOFFMAN and E. W. MARKOW

Federal Telephone and Radio Corporation, Newark, New Jersey

RECENT YEARS have witnessed increasing interest in radio communication to moving vehicles, particularly in the emergency services for police and fire protection. Its convenience for dispatching taxicabs, public-utility repair crews, and similar services has not been overlooked. A description is given of transmitting and receiving equipment for mobile, main, and remote stations. Operation is in the 30- to 44-megacycle band using frequency modulation. When Selecto Call is employed, up to four mobile stations or groups of stations may be called individually, without disturbing unwanted mobile stations.

• • •

A line of equipment has been designed to provide single-frequency two-way radio communication between a land[1] station and any number of mobile stations.

The equipment operates in the band from 30 to 44 megacycles per second, assigned by the Federal Communications Commission for mobile communication services, and uses frequency modulation.

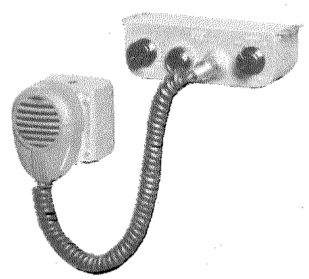
The complete system consists of equipment for mobile and for both central and remote land stations. It is characterized by simplicity of operation, small size of mobile equipment, improved receiver squelch circuit, low power drain for mobile equipment during stand-by, plug-in arrangement of units for ease of replacement and servicing, and a self-contained selective calling system. This selective calling feature, or "Selecto Call," permits the land station to call any of four mobile stations or groups of stations without alerting other mobile units in the system.

In some instances, incorporation of Selecto Call may avoid duplication of land-station equipments. For example, a small community may combine its mobile communication with fire, police, and ambulance services by using a single headquarters station and equipping cars of the separate departments with different Selecto Call decoders. In this way, cars of any single department could be alerted without disturbing any other cars.

Mobile receivers are available without the Selecto Call decoder unit which is plugged in if required. Similarly, the land-station equipment need not include the special signaling devices for Selecto Call if this feature is not needed.

1. *Mobile Equipment*

The standard mobile equipment, shown in Fig. 1, consists of a 25- to 50-watt transmitter, receiver, control box, loudspeaker, microphone,



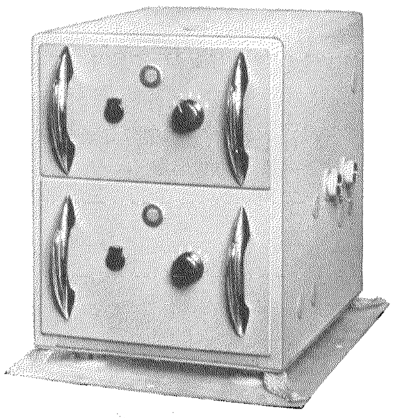


Fig. 1—The mobile receiver and transmitter are mounted in the cabinet and may be placed in any convenient space in the automobile. The microphone and control box are usually mounted on the instrument board.

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[1] A "land" station is a radiotelephone installation in a fixed location. The older term of "fixed" station commonly denotes one in point-to-point service.

microphone hook, and whip antenna. Both the receiver and transmitter are on separate plug-in chassis, which are inserted in either a vertical or horizontal housing. The control box is normally mounted on the instrument panel of the automobile.

The transmitter and receiver units predominantly employ miniature tubes. Crystal-controlled oscillators keep both receiver and transmitter on frequency. The over-all response-frequency characteristic of the system is shown in Fig. 2. The equipment can be tuned to operate at any radio frequency between 30 and 44 megacycles. Power is obtained from the 6-volt automobile storage battery. The current drain during stand-by periods is 7.12 amperes; while transmitting, the current increases to 33.2 and 50.5 amperes for the 25- and 50-watt transmitters, respectively.

Operation of the mobile equipment requires only lifting the microphone off the hook, depressing the microphone switch, and speaking. The receiver volume is adjusted from the control box to suit conditions of ambient acoustical noise which varies with speed and traffic conditions, among other things. The receiver may be operated in any one of three conditions designated as carrier, monitor, and Selecto Call.

In the "carrier" position, the receiver audio-frequency channel is silenced by a squelch circuit which maintains control until an on-frequency carrier is received. This completely removes all annoying electrical noises, which would normally be present in the absence of a carrier. The receiver is completely operative only when there is a transmission on the frequency to which it is tuned. The squelch circuit is of high sensitivity, turning on the audio-frequency system with signals producing only 6 decibels of noise quieting. Such signals are too weak to be understood with ease, therefore, no readable signal will be undetected when the receiver is in "carrier" position.

In the "monitor" position, the squelch circuit is inoperative and the audio-frequency channel is not silenced. The operator hears the annoying hissing and rushing noise characteristic of all high-gain receivers in the absence of a quieting signal. This position is useful for the reception of very weak and fading signals and to check the operation of the receiver.

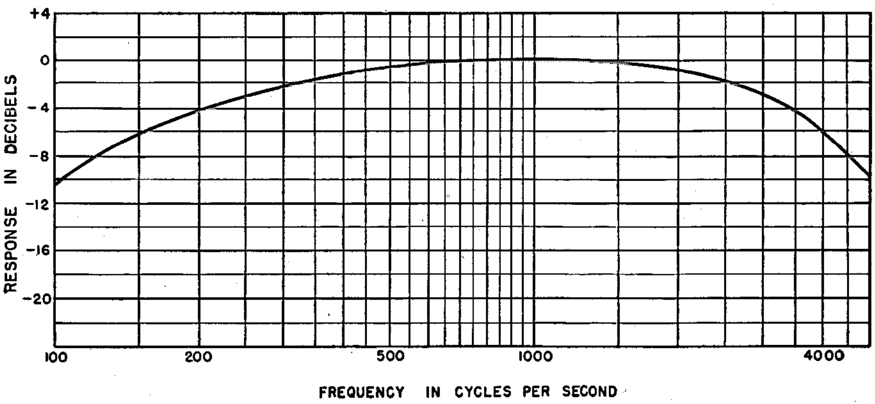

Fig. 2—Response-frequency characteristic of the mobile equipment. Reference level is the response at 1000 cycles.

In the "Selecto Call" position, a frequency-selective decoding circuit renders the receiver audio-frequency system operative only in response to a predetermined calling signal. When the microphone is lifted from its hanger, the receiver is switched to the "carrier" position. It is unnecessary thereafter for the land station to repeat the Selecto Call.

2. *Transmitter*

The mobile transmitter, a block diagram of which is shown in Fig. 3, consists essentially of a crystal-controlled master oscillator, a phase-shift variable-transconductance modulator, and four frequency-multiplying amplifiers which drive the output stage. The final amplifier consists of either one type 5516 tube or two in parallel to provide, respectively, 25 or 50 watts of radio-frequency power to the antenna.

The oscillator is of the conventional crystal-controlled type employing a 6AK5. Frequency shift is less than ±0.01 percent over a temperature range from -30 to $+70$ degrees centigrade (-22 to $+158$ degrees Fahrenheit). After modulation, the signal is applied to a series of frequency multipliers which operate as saturated class-C amplifiers. The output of the power amplifier stage, also operating class C, is coupled to the antenna through a tuned π network. The transmitter may be operated at any frequency between 30 and 44 megacycles by inserting a crystal of the proper frequency and adjusting the tuned circuits.

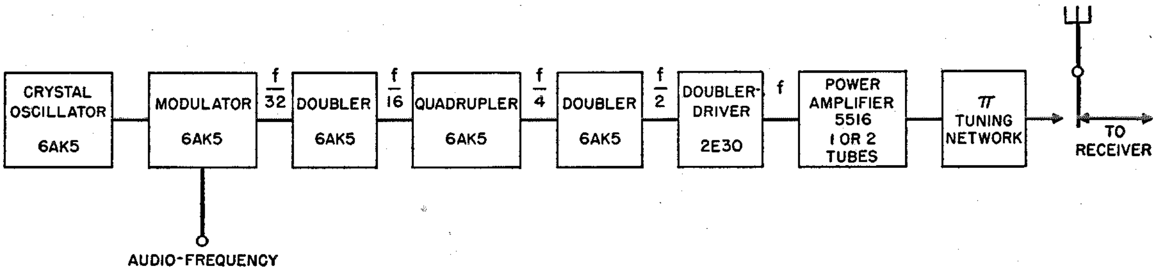

Fig. 3—Schematic diagram of mobile transmitter. Only one power-amplifier tube is used for the 25-watt rating and two for 50 watts. Closing the microphone switch operates the antenna change-over relay and another relay in the primary circuit of the transmitter dynamotor.

The variable-transconductance phase-shift modulator is shown in Fig. 4. A 6AK5 is connected as a triode. A very high resistance at R_k produces a large grid bias with a consequent low tube transconductance. The master-oscillator voltage E_g applies two voltages to the tuned plate circuit Z_p of the modulator: one through C_{gp} the grid-to-plate capacitance, and the other E_{pg_m} through the tube conductance. Thus the voltage developed across the tuned plate circuit Z_p consists of two components, nearly opposite in phase with respect to each other. With no modulating signal E_a these two components are equal in magnitude, and the resultant E_{po} is shown in Fig. 5. When an audio-frequency signal E_a is applied from the microphone to the grid, the instantaneous value of tube transconductance and of E_{pg_m} varies in accordance with the instantaneous magnitude of the audio-frequency voltage as shown by two arrow heads on the vector diagram. Thus as the audio-frequency voltage increases and decreases through its cycle, the component across the tuned circuit Z_p resulting from tube conductance changes in magnitude. However, the component caused by the grid-to-plate capacitance E_{pc} remains constant, and the resultant radio-frequency voltage across the tuned plate circuit shifts in phase through the angle α. Thus, instantaneous changes in the magnitude of the audio-frequency signal cause

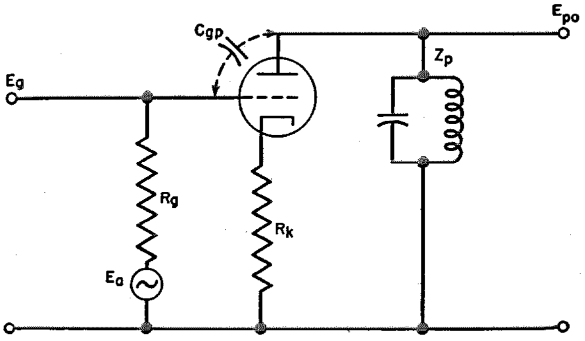

Fig. 4—Variable-transconductance phase-shift modulator.

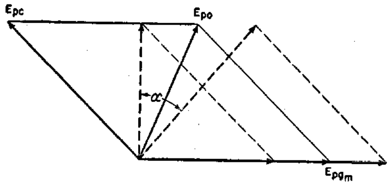

Fig. 5—Vector analysis of the operation of the circuit of Fig. 4. E_{pc} and E_{pg_m} are the voltages on Z_p as a result of E_g being applied through C_{gp} and through the conductance of the tube, respectively. A voltage E_a applied to the tube grid, changes the magnitude of E_{pg_m} without affecting E_{pc}, producing a phase variation α in the resultant output radio-frequency voltage E_{po}.

corresponding excursions of the phase angle and produce phase deviations of the radio-frequency voltage from the phase obtained in the absence of modulation. The frequency of the modulating signal expresses itself as the rate with which the phase of the resultant radio-frequency voltage changes. The highest audio-frequency modulating signal used is about 3500 cycles. The maximum equivalent frequency deviation does not exceed 15 kilocycles and complies with the regulations of the Federal Communications Commission. Any amplitude modulation introduced is eliminated by the chain of saturated radio-frequency amplifier stages.

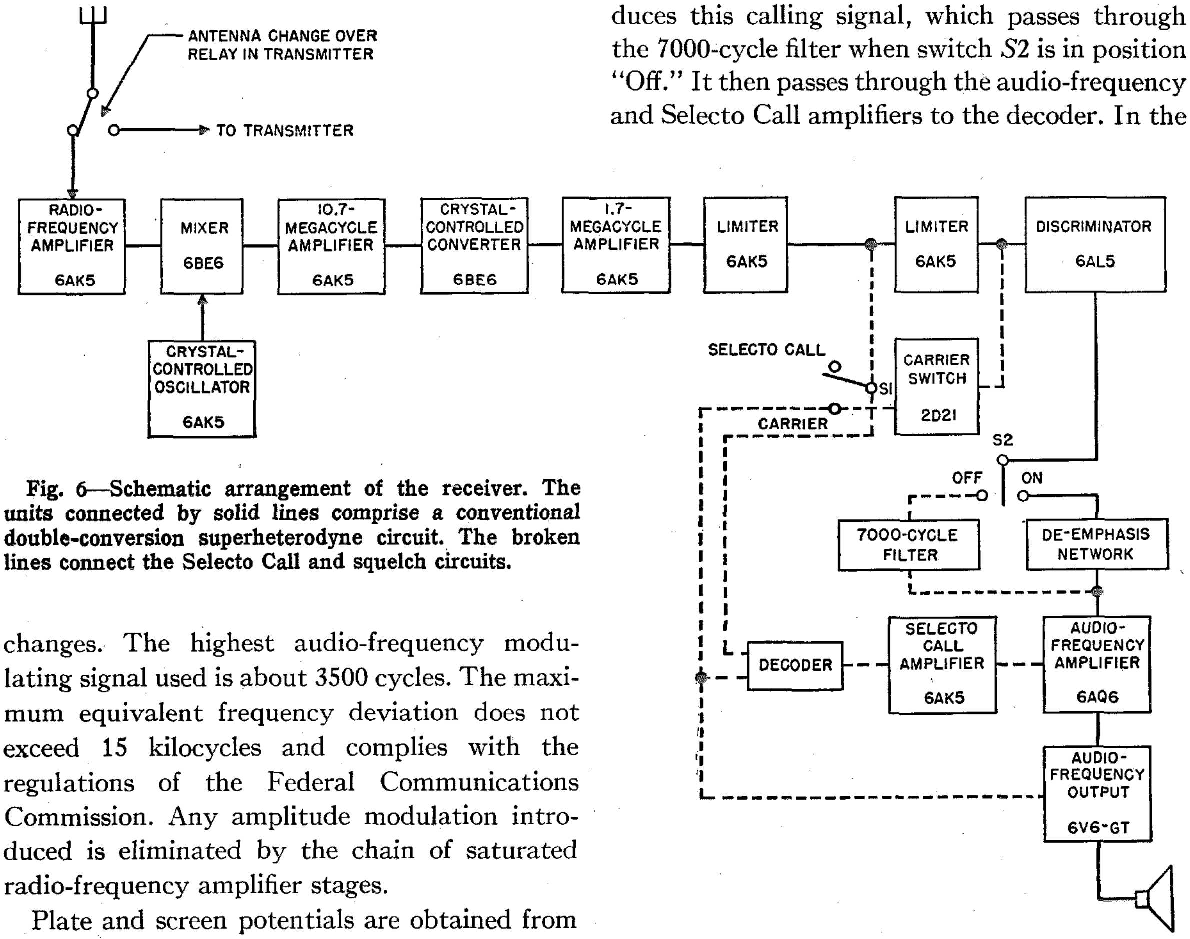
ANTENNA CHANGE OVER RELAY IN TRANSMITTER
TO TRANSMITTER
RADIO-FREQUENCY AMPLIFIER 6AK5
MIXER 6BE6
10.7-MEGACYCLE AMPLIFIER 6AK5
CRYSTAL-CONTROLLED CONVERTER 6BE6
1.7-MEGACYCLE AMPLIFIER 6AK5
LIMITER 6AK5
LIMITER 6AK5
DISCRIMINATOR 6AL5
CRYSTAL-CONTROLLED OSCILLATOR 6AK5
SELECTO CALL
S1
CARRIER
CARRIER SWITCH 2D21
S2
OFF
ON
7000-CYCLE FILTER
DE-EMPHASIS NETWORK
DECODER
SELECTO CALL AMPLIFIER 6AK5
AUDIO-FREQUENCY AMPLIFIER 6AQ6
AUDIO-FREQUENCY OUTPUT 6V6-GT

Fig. 6—Schematic arrangement of the receiver. The units connected by solid lines comprise a conventional double-conversion superheterodyne circuit. The broken lines connect the Selecto Call and squelch circuits.

Plate and screen potentials are obtained from a built-in dynamotor operated from the car battery. For land installations, the dynamotor is replaced by a power pack operated from the 117-volt 50/60 cycle supply lines.

3. Receiver

A block diagram of the receiver is shown in Fig. 6. The units connected by solid lines comprise a conventional double-conversion superheterodyne. The components connected by broken lines comprise the Selecto Call mechanism. When the switch *S*2 is in the "On" position, and switch *S*1 is in the "Carrier" position, the receiver is operative without being under the control of the Selecto Call mechanism.

The Selecto Call actuating signal is a single low-frequency wave between 150 and 442 cycles, which modulates a 7000-cycle subcarrier. This combined signal modulates the transmitter carrier. The discriminator in the receiver reproduces this calling signal, which passes through the 7000-cycle filter when switch *S*2 is in position "Off." It then passes through the audio-frequency and Selecto Call amplifiers to the decoder. In the process, the 7000-cycle subcarrier is discarded and, if the lower-frequency modulation is of the proper frequency, the decoder operates. The decoder is essentially a single-pole single-throw switch, which is normally in the open position. When the decoder is actuated, the switch is closed and the audio-frequency system becomes operative.

If switch *S*2 is thrown to the "On" position, the audio-frequency channel will continue operative as long as the radio-frequency carrier is being received. If the carrier is removed, the audio-frequency system is disabled and the

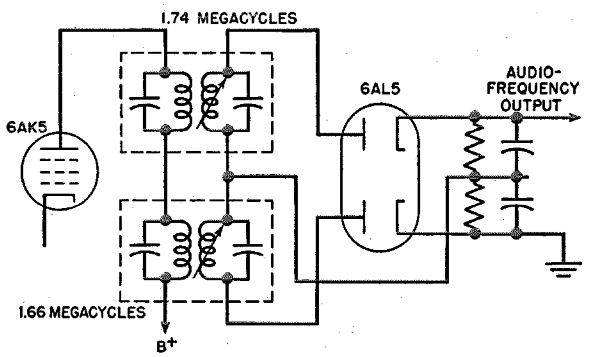

Fig. 7—Discriminator circuit. The capacitor in the primary of the lower transformer tuned to 1.66 megacycles is adjusted to produce zero direct-current output for an input signal of 1.7 megacycles.

decoding signal will be required to open the channel again. A carrier alone, or a carrier with any modulation other than that for which the decoder was designed, will not operate the receiver.

Switch *S*2 is shifted from the "Off" to the "On" position by lifting the microphone from its hook.

Decoders are available for four frequencies which may be assigned to individual or groups of receivers to meet the requirements of various installations.

Modulation of the transmitter carrier with frequencies between 150 and 442 cycles results in very small deviations of frequency with the phase modulator employed. The use of a subcarrier of 7000 cycles, which is modulated in turn by the low-frequency decoding signal, avoids this difficulty and, by giving a full 15-kilocycle deviation, provides a maximum signal to the receiver. Only the frequencies between 150 and 442 cycles reach the decoder coil, the 7000-cycle frequency being attenuated in a low-pass filter in the Selecto Call amplifier.

3.1 Squelch Circuit

The squelch circuit employs a type 2D21 thyratron, which conducts current in the stand-by condition. A relay in its plate circuit is operated and disconnects the voltage supply to the screen of the 6V6-GT audio-frequency output tube, rendering the audio-frequency channel inoperative.

The grid of the thyratron receives voltages from two sources. A negative voltage is taken from the bias resistor in the grid circuit of the second limiter tube. The second voltage is obtained through a high-pass filter from the plate of the second limiter tube. Under stand-by conditions, the limiter is noise-saturated and the positive peaks from the plate circuit exceed the steady negative voltage derived from the grid circuit. The positive noise pulses dominate and are strong enough to fire the thyratron. Resistance and capacitance in the grid circuit of the thyratron produce relatively long operating periods. At the end of each operation, another noise pulse triggers the thyratron and sufficient plate current is maintained to hold the relay in the operated position.

If a signal is received at the frequency to which the receiver is tuned, there will be no change in the limiter grid voltage. However, the positive peaks of noise voltage are reduced and cannot maintain the thyratron in operation. The plate relay then opens and voltage is applied to the screen of the audio-frequency output tube, placing the output channel in operation.

If an adjacent-channel signal of sufficient strength is present, the early radio- and intermediate-frequency stages tend to saturate and reduce the noise signal on the limiter grids. With sufficient input signal, the limiter grid bias will drop to zero, in the absence of noise pulses from the early stages, as will the discriminator output voltage in the absence of noise. The discriminator output voltage also is applied to the grid of the thyratron through suitable isolating filters and the thyratron bias becoming zero, the tube remains fired. This effect protects the receiver against adjacent-channel input signals of all signal levels.

3.2 Crystal Oscillator

The first local oscillator uses a crystal which oscillates directly at the third harmonic of its fundamental frequency. No energy is present at the fundamental frequency. Higher output and greater frequency stability are obtained than from the conventional mode of operation.

3.3 DISCRIMINATOR

The discriminator circuit, shown in Fig. 7, is a double-tuned dual-transformer type. The transformers are tuned to peaks 80 kilocycles apart (1.7 megacycles ±40 kilocycles), and the outputs are phased so that the linear portion of the pass-band curve between the transformer peaks can be utilized as a frequency-modulation discriminator. Adjusting the primary inductance of the lower-frequency transformer will vary the amplitude of the signal in the secondary winding and permit center-frequency zero adjustment.

The over-all sensitivity is such that a 0.5-microvolt radio-frequency signal will produce 20 decibels of thermal-noise quieting, when measured in accordance with the procedure recommended by the Radio Manufacturers Association.

The audio-frequency system utilizes a 6V6-GT output tube driven by a 6AQ6 amplifier. A maximum output of 1.5 watts is obtained, and about 1 watt with not more than 10-percent harmonic distortion.

The over-all battery drain for the receiver in mobile service is about 6 amperes, stand-by, and 7 amperes in full operation. A power supply, employing an OZ4A gas rectifier tube is used for 6-volt operation. This is replaced by a conventional full-wave rectifier using a 5Y3-GT tube when in land-station service.

4. Land-Station Console

The land-station console, shown in Fig. 8, is a complete self-contained equipment which will fit on top of a standard office desk. It is designed specifically for operation with the mobile radio equipment already described, and contains the special Selecto Call oscillator-modulator circuits. The equipment includes a 50-watt transmitter, power supply, speech amplifier, and receiver. A second receiver may be included for monitoring. All units are of the plug-in type. The sloping front panel minimizes glare on the meter and clock windows. Chrome-plated trim and a baked gray wrinkle finish make for attractive appearance.

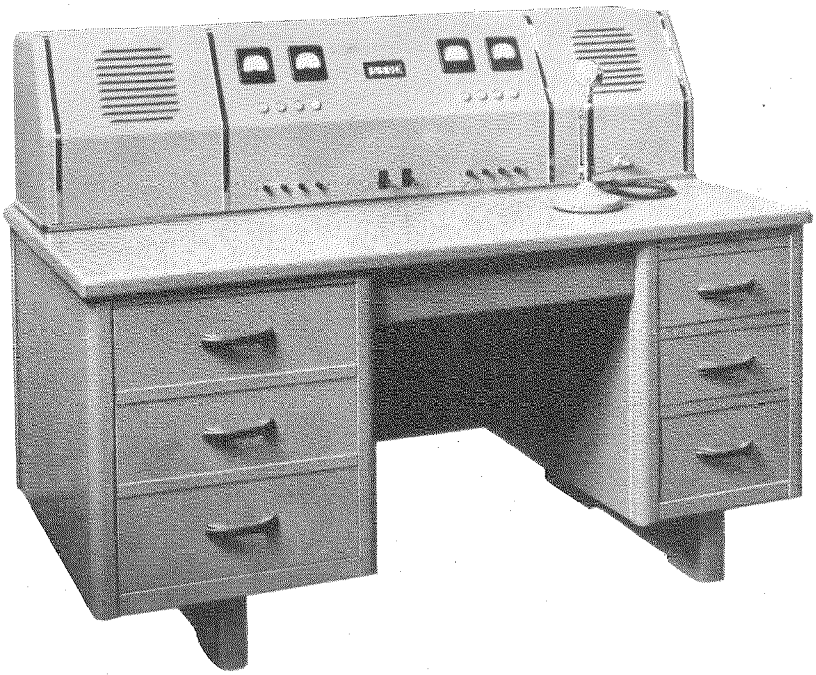

Fig. 8—Land-station console containing 50-watt transmitter, power supply, speech amplifier, receiver, and monitoring receiver.

The transmitter and receiver are interchangeable with the corresponding units of the mobile equipment, except that power supplies are changed to permit operation from 117-volt 50/60-cycle mains. The receiver is not equipped with the Selecto Call decoders and will, therefore, respond to all mobile transmissions. Fig. 9 is a rear view, which discloses how these standard units are mounted in the console.

When calling a mobile receiver that is using Selecto Call, the land-station operator must first select, by depressing a push button, the decoder frequency to which the mobile receiver will respond. To call several groups of cars, the station operator may depress the desired push buttons in any order.

Operation of a lever switch automatically transmits the Selecto Call signal for about one second. The transmitter remains on as long as this switch is operated. A panel light indicates the completion of the calling signal and that the equipment is ready for the attention tone or speech. Depressing the attention-tone lever, transmits an 800-cycle note.

Gripping the microphone handle permits voice modulation if neither lever switch is depressed. The microphone switch also turns on the transmitter and permits immediate speech and rapid back-and-forth communication with the mobile unit. If the lever switches are depressed simultaneously and the operator starts to speak or transmit attention tone before the completion of the calling signal, the calling signal will take precedence over the other signals. Likewise, attention tone takes precedence over speech.

4.1 Transmitter Power Supply

The transmitter power supply operates from the 117-volt alternating-current mains and supplies heater, relay, and plate power to the transmitter. The 6-volt relay power is obtained from a bridge-type selenium rectifier and filter. Plate power is obtained from a conventional full-wave rectifier circuit using a 5R4-GY vacuum tube. Heater power is taken from the secondary winding of the filament transformer.

4.2 Speech Amplifier

The speech amplifier includes oscillators for four Selecto Call signals, an 800-cycle attention tone, and a 7000-cycle subcarrier. It also contains a modulator, a two-stage preamplifier, and a modulation level limiter. The Selecto Call oscillators generate the low-frequency signal to which the receiver decoders are responsive. The 7000-cycle oscillator provides the carrier for the low-frequency signal; their combination occurs in the modulator stage. The timing of the Selecto tone transmission is controlled by a large capacitor in series with a relay; the relay opens and breaks the Selecto oscillator circuits after an interval of about one second.

4.3 Preamplifier

The preamplifier provides gain for the crystal microphone and consists of one stage of audio-frequency amplification, one limiter stage, and one power output stage, employing 6SL7-GT, 6AL5, and 6V6-GT tubes, respectively. A conventional resistance-coupled circuit is used.

The function of the limiter is to provide constant amplifier output to keep the transmitter frequency deviations within the required limits of ±15 kilocycles. The circuit is arranged so that one-half of the 6AL5 limits the positive peaks of modulating voltage and the other half controls the negative peaks. The limiter is capable of maintaining output variations to with-

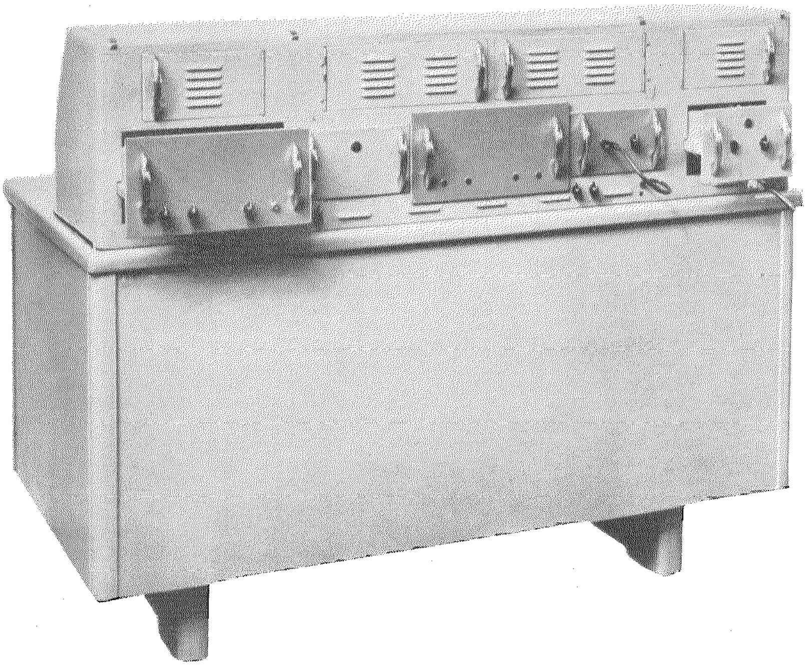

Fig. 9—Rear view of land-station console showing how the component units are mounted.

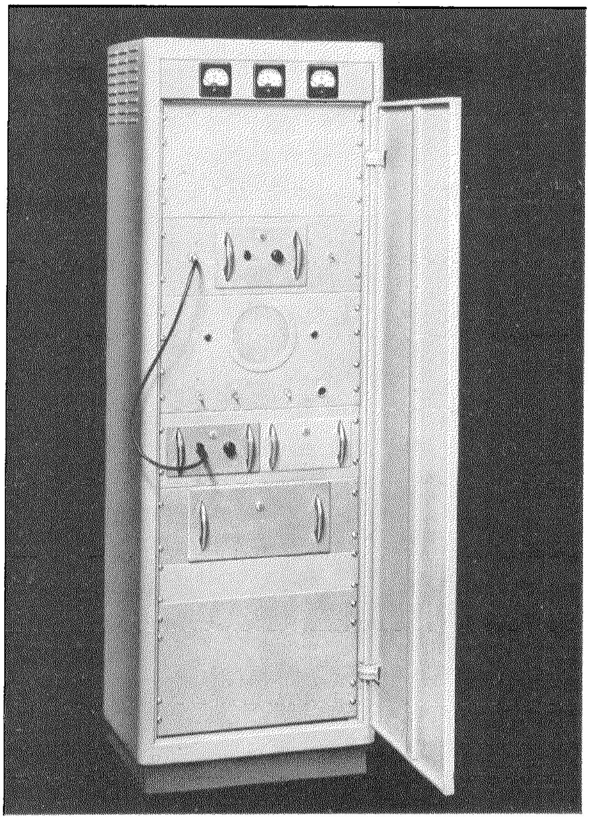

Fig. 10—Remotely controlled land-station 250-watt transmitter and receiver, front door open.

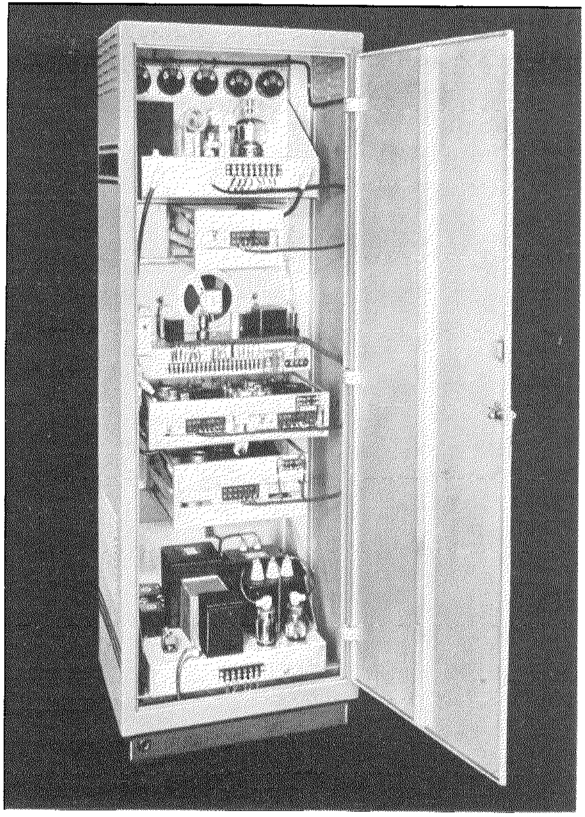

Fig. 12—Rear view of 250-watt remotely controlled equipment, with door open.

in 10 decibels for a 40-decibel variation of input signal. This provides constant output for all normal microphone uses. The level at which limiting begins, as well as the output of the amplifier, may be adjusted by potentiometers at the rear of the console.

4.4 Selecto Call Oscillators

Each Selecto Call oscillator is a two-tube Wien bridge. One section of a 6SL7-GT and of a 6SN7-GT, both dual triodes, is used for each oscillator. Components have been chosen to provide optimum frequency and amplitude stability. The subcarrier and attention tone are also generated by similar Wien bridge oscillators.

4.5 Modulator

The modulator stage consists of a dual-triode 6SN7-GT connected as a conventional Heising amplitude modulator. The modulator is very

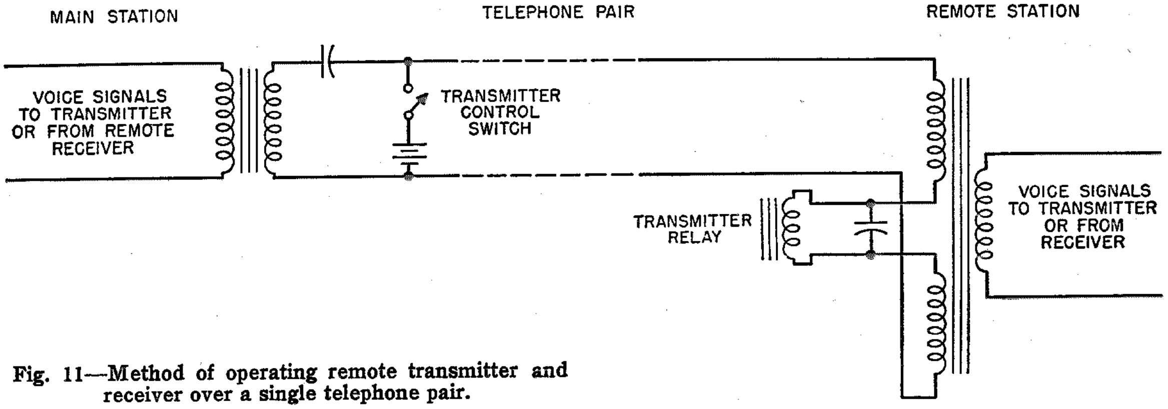
MAIN STATION
TELEPHONE PAIR
REMOTE STATION
VOICE SIGNALS TO TRANSMITTER OR FROM REMOTE RECEIVER
TRANSMITTER CONTROL SWITCH
TRANSMITTER RELAY
VOICE SIGNALS TO TRANSMITTER OR FROM RECEIVER

Fig. 11—Method of operating remote transmitter and receiver over a single telephone pair.

simple and produces practically 100-percent modulation of the 7000-cycle subcarrier by the low-frequency Selecto signal. A tank circuit resonant at 7000 cycles is shunted across the output to pass the subcarriers but not the low-frequency signals.

5. *Remote-Station Equipment*

It is not always possible for a single land station of reasonable power to cover completely the area which must be protected by an emergency service. Consequently, remotely located transmitters and receivers are necessary in some installations where the area to be covered is extensive or where highly industrialized regions cause abnormal attenuation of radio waves and high noise levels, both radio and acoustic.

The equipment, shown in Fig. 10, includes a 50-watt transmitter, one or two receivers, a line-terminating unit, and a transmitter power supply. The transmitter and receiver units are the same as those used in mobile service. This apparatus may be remotely controlled over a telephone line. It operates from a 117-volt, 50/60-cycle source.

The control console is similar in appearance and manner of operation to the main-station console. It provides all the controls and facilities required for operating the remote transmitting and receiving station from a location as far as 20 miles from the main station, and includes a speech amplifier and remote-control unit.

The speech amplifier is identical to the unit in the main-station console; the audio-frequency output is connected to a 600-ohm telephone line. The signals received over the telephone line from the remote receiver pass through the amplifier. The Selecto Call, attention tone, and other circuits operate as previously described.

The remote-control relay unit includes power-level controls for the telephone line, microphone and loudspeaker, and control relays, as well as matching networks necessary to permit remote operation of the transmitter and receiver units. Both audio-frequency and control signals are transmitted over a single line as shown in Fig. 11.

The transmitter-receiver cabinet is arranged to permit receivers tuned to different frequencies to be used over the same telephone line so one operator may monitor several stations. The circuit is arranged so that in the event of simultaneous reception through both receivers, the guard or monitor receiver is automatically silenced.

6. *Higher-Power Transmitter*

A remotely controlled land station of higher power is shown in Fig. 12. It employs a 250-watt power amplifier and a higher-voltage plate supply. The 25-watt transmitter functions as an exciter for the power amplifier, which employs a 4-250 vacuum tube in a single-ended stage. Tuning is accomplished from the front panel. A current overload relay removes plate voltage to avoid damage to tubes and components in the event of any failure. A blower, thermostatically controlled, provides positive ventilation. The cabinets for both remote stations have individually locked doors, the rear door being electrically interlocked to avoid danger from high voltages.

7. *152—162-Megacycle Equipment*

In addition to the 30–44-megacycle equipment described above, there has been developed similar apparatus for operation in the 152–162-megacycle band. The equipments differ only in circuit details; over-all sizes, methods of operations, and available equipments are identical.

The 30–44-megacycle equipments have their major application in rural services where large relatively noise-free areas are to be covered. The 152–162-megacycle equipments, although not providing quite as great a distance range, are more suitable for congested urban areas where ambient noise may be high. Measurements have shown that static and man-made radio noise is much less at the higher frequencies as compared with 30–44 megacycles, so that coverage will probably be better using 152–162 megacycles in such noisy areas because of the improved signal-to-noise ratio.

Medium-Power Triode for 600 Megacycles*

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THE air-cooled triode, L600E, was developed to produce a peak pulse power of 25 kilowatts at 600 megacycles for radar operation. The water-cooled 6C22 was then developed to provide higher efficiency in continuous-wave operation at that frequency. The construction and characteristics of these tubes are described. In an experimental crystal-controlled transmitter, the 6C22 has delivered 500 watts at 600 megacycles.

• • •

The large peak power required by pulse radar systems demands vacuum tubes capable of delivering high peak emission currents at high voltages. The tubes are usually employed as oscillators and the modulation, in the form of the pulses, is applied to the anode. The pulse is, thus, the envelope of several radio-frequency cycles. The radio frequency is determined by the oscillator circuit constants.

The high instantaneous voltages that exist during operation make it desirable to avoid spacing insulators between tube electrodes and to lengthen the external glass paths as much as possible. At the same time, rigidity of the electrode structure must be attained.

Interchangeability of tubes in any equipment is necessary and applies not only to the electrical characteristics, but to the physical mounting of the tube in the circuit as well.

In addition, the tubes must maintain their original characteristics and the available thermionic emission must not decay appreciably over long periods of operation. Satisfactory life requirements have been met in pulse radar tubes employing either thoriated-tungsten filaments or indirectly heated oxide-coated unipotential cathodes.

1. L600E

As radar development proceeded, it was evident that higher frequencies and powers had much to offer in increasing accuracy and range. Consequently, the development of a tube capable of delivering peak pulse powers of 25 kilowatts at 600 megacycles was undertaken in the latter part of 1941. After several designs had been made and tested, a developmental tube known as the L600E proved most promising. Technical information on this tube is given in Table I.

TABLE I

CHARACTERISTICS OF L600E

Filament	Thoriated Tungsten
Filament volts	6
Filament amperes	13.5
Amplification factor	20
Mutual conductance (ma/volt, $I_b=2a$, $E_c=-100v$)	10
Maximum anode volts (pulse)	25,000
Maximum anode dissipation (kilowatts)	0.3
Capacitance ($\mu\mu f$) C_{gp}	4
C_{gf}	5
C_{pf}	0.25

The L600E consists of a bifilar thoriated-tungsten-filament emitter with its electrical center brought out to a pin terminal, a squirrel-cage-type grid, and a re-entrant copper anode with a bayonet ring at its lower end to provide a simple means for securing and positioning the tube in the socket.

With anode modulation in a concentric-line-type oscillator circuit, 25 kilowatts of peak power output could be produced at 600 megacycles by a single tube. In one application using grid modulation, 7200 anode volts, and 1.62 pulses per second, each of 200 microseconds duration, 7.3 kilowatts of peak power were developed at 600 megacycles and 33 percent efficiency.

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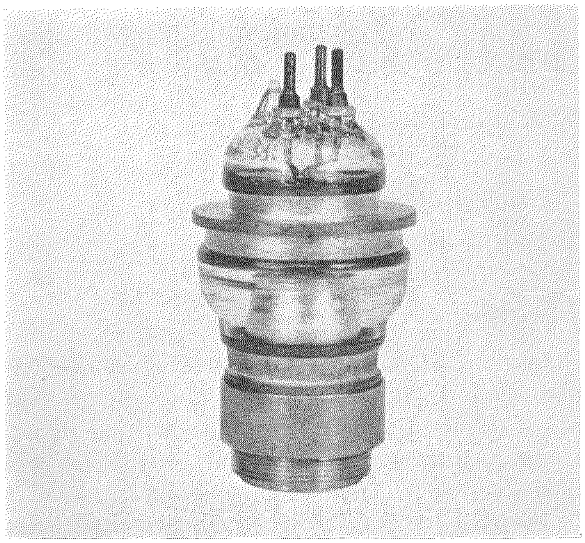

Fig. 1—6C22.

In high-frequency tubes, transit-time effects must be seriously considered. By decreasing interelectrode clearances, the deleterious effects of the long time of electron travel compared with the period of the applied potentials are reduced. Transit-time effects are also decreased when high anode voltages are employed. Thus, a tube that may operate at a given high frequency with good efficiency under pulse conditions where high voltages are employed, may exhibit very low efficiency in continuous-wave operation at the same frequency because lower anode voltages must be employed.

One result of transit-time effects is that not all electrons emitted by the filament or cathode when the grid is positive reach the anode. Many emitted electrons return to the cathode space-charge region. This causes a decrease in anode current and power output. To increase the anode current, the filament or cathode must supply a larger quantity of electrons to compensate for those electrons which do not reach the anode. It is obvious that the emission capabilities of the filament or cathode must be much greater in high-frequency applications than in low-frequency applications. The emission required may be several times as great as in low-frequency applications where transit-time effects are negligible.

2. *6C22*

While the L600E performed well as a pulse tube, it was unsatisfactory for continuous-wave operation. Employed as a continuous-wave oscillator in a concentric-line grid-separation circuit with 1000 anode volts, the efficiency at 600 megacycles was only 10 percent, although at 300 megacycles the efficiency was 40 percent. Consequently, a design formerly known as the L600N, now designated as the 6C22, was developed.

2.1 Construction

With the exception of the filament terminals and screw-type water jacket on the anode, the 6C22 is similar in outward appearance to the L600E. It is shown with the jacket in Fig. 1 and a sectional view appears in Fig. 2.

Internally, it differs dimensionally from the L600E. The filament-wire diameter is 25 percent greater and the number of turns has been increased by approximately 35 percent. The grid-filament and grid-anode clearances were reduced by approximately 60 percent.

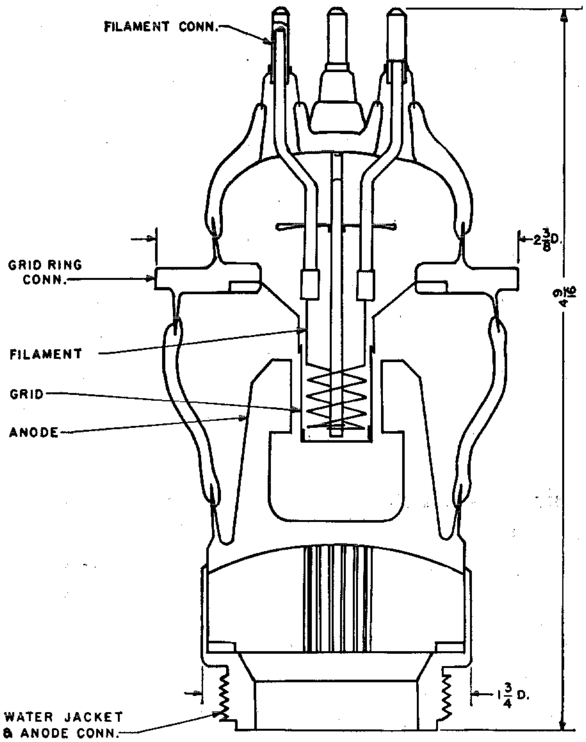

Fig. 2—Cross-sectional view of 6C22.

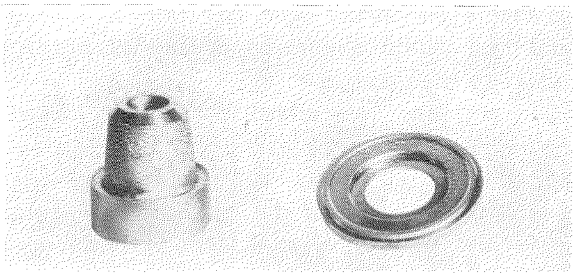

Fig. 3—Re-entrant anode and grid ring of 6C22.

The result of this reduction in clearances and increased filament surface area was a much higher perveance and a lower amplification factor. The additional surface area increased the available thermionic emission.

2.2 Filament

The helical bifilar thoriated-tungsten filament is made of 0.025-inch-diameter wire and has an active emissive surface area of 3 square centimeters. The filament ends and center lead are terminated in monel-kovar cup assemblies sealed to a molded glass flare. This construction has the advantage of great mechanical strength and resistance to impact at the seals.

A shadowgraph comparator with a magnification of 10 times is used to inspect the filament for alignment of the turns to minimize unequal grid heating.

2.3 Grid

The grid is of the squirrel-cage type, consisting of 32 wires, 0.008 inch in diameter, spot welded to and supported by a low-inductance cone. After completion, the grid is cold stretched on the welding mandrel and then hot stretched in vacuum on a special fixture to relieve all stresses and equalize the tension in the cage wires. In addition, this procedure assures equal heating and expansion of the grid in operation. The grid is inspected on the shadowgraph comparator for wire straightness.

2.4 Grid Ring and Anode

The grid ring and the anode are made of oxygen-free high-conductivity copper and lend themselves to fabrication by coining because of their unusual shape. The ring is coined from a washer $2\frac{1}{4}$ inches in outside diameter, 1 inch in inside diameter, and $\frac{3}{16}$ inch thick. Under 250 tons pressure it assumes the shape shown in Fig. 3. The stubs are then trimmed so that they are $\frac{1}{16}$ inch square, and the feather edges are formed from these stubs. The re-entrant anode is coined from $1\frac{5}{16}$-inch-diameter bar stock 2 inches long under 300 tons pressure to the shape shown in Fig. 3 and is then slotted on the water-jacket end. The feather edge is formed in the same manner as that of the grid ring in a conventional machine lathe especially adapted for the purpose. The feather edges are carefully inspected for dimensions and flaws, after which the parts are cleaned and then glassed.

2.5 Assembly Procedure

The present bifilar filament is mounted on the tungsten leads of the molded flare, as shown in Fig. 4, and carburized. This assembly has the prebeaded grid ring sealed to it. Special fixtures

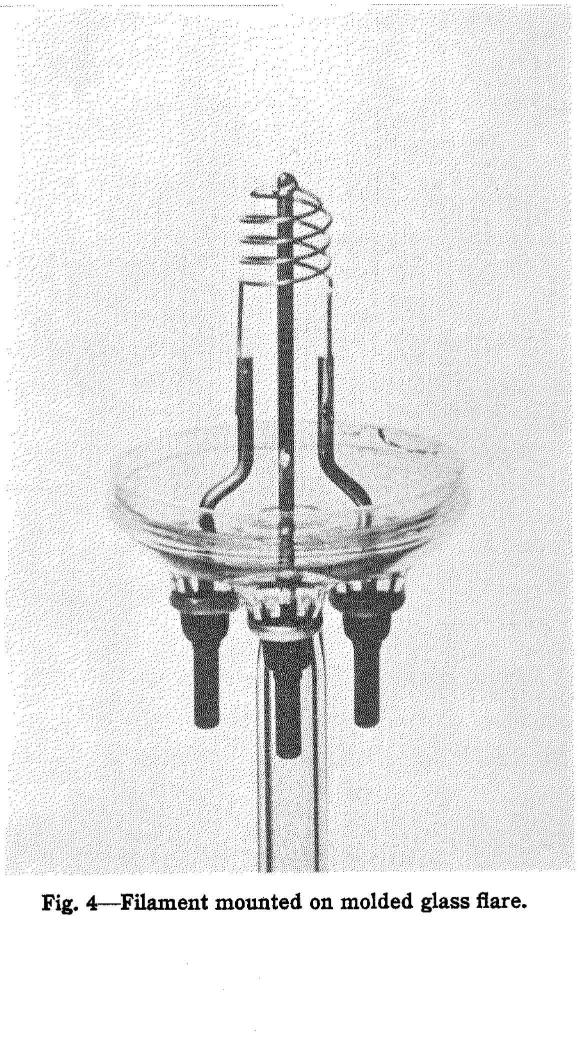

Fig. 4—Filament mounted on molded glass flare.

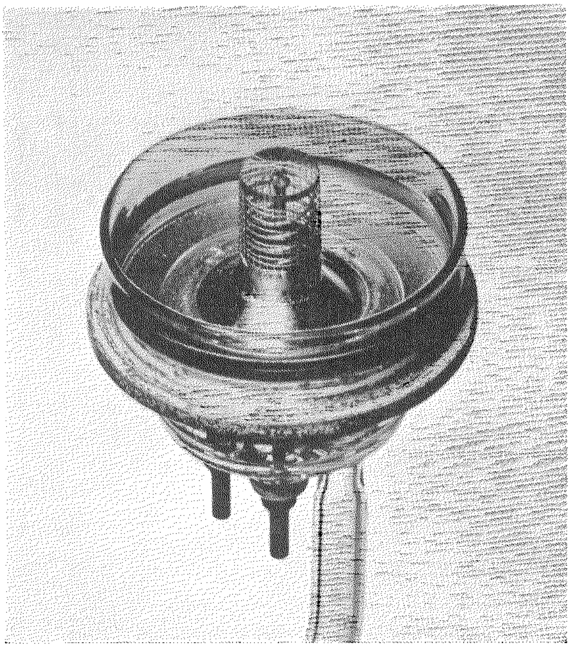

Fig. 5—Grid and filament assembly. The grid connecting ring and cone support may be clearly seen.

are employed during this operation to assure axial and radial alignment of this ring with respect to the filament.

The grid is next assembled to its ring by means of a copper clamp ring and four flat-head screws. Before these screws are tightened, the grid is carefully aligned with the filament by again employing the shadowgraph. The resultant assembly is shown in Fig. 5.

The anode, with glass of proper length and shape already sealed to its feather edge, is now joined to the glass skirt on the grid connecting ring after careful alignment with respect to each other. This operation is shown in Fig. 6.

During all sealing operations, an atmosphere of nitrogen gas is employed to prevent oxidation of the tube parts.

2.6 CHARACTERISTICS

As a result of the care used in making and assembling the electrodes, the electrical char-

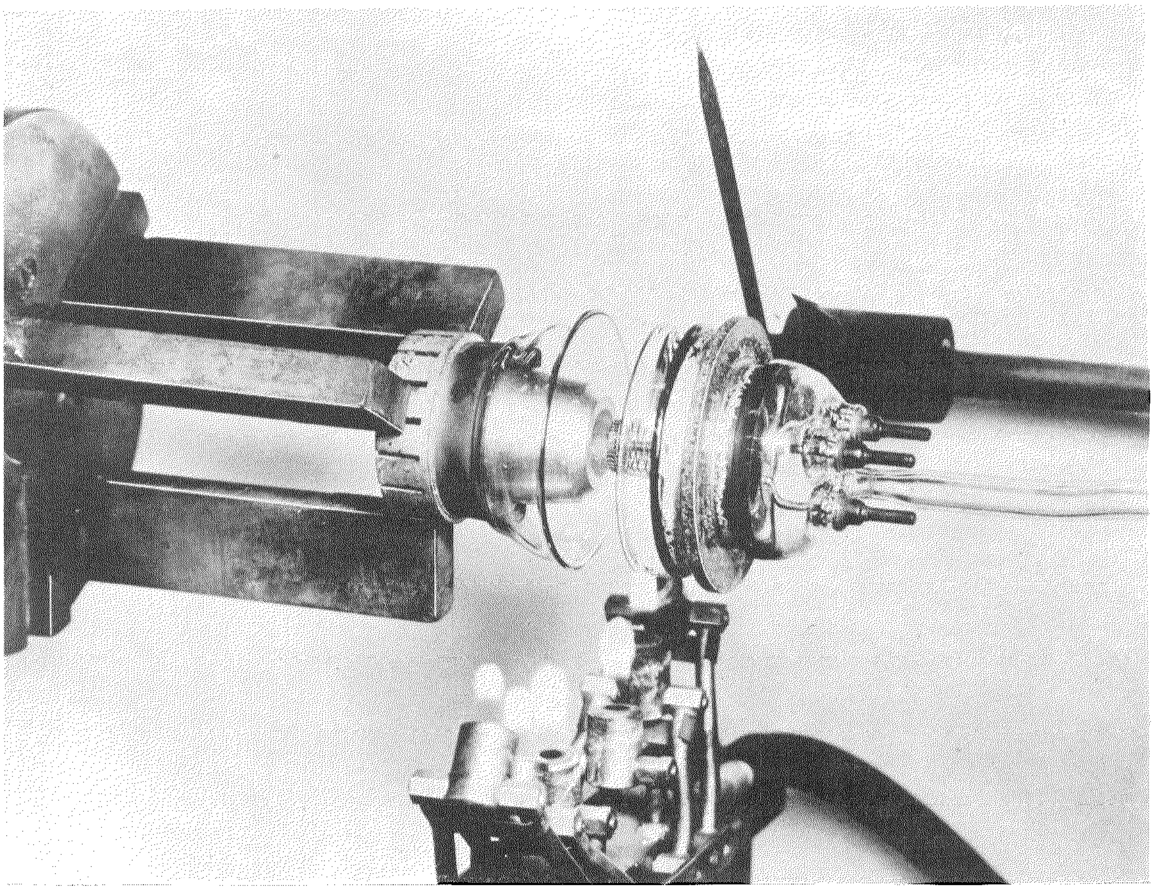

Fig. 6—The glass-mounted anode, held by the jaws of the lathe chuck, and the filament-grid assembly are joined at the two glass surfaces.

TABLE II

Characteristics of 6C22

Filament	Thoriated Tungsten
Filament volts	6.5
Filament amperes	18
Amplification factor	9
Mutual conductance (ma/volt, $I_b=3a$, $E_c=-100v$)	13
Maximum anode volts	3000
Maximum anode dissipation (kilowatts)	2
Maximum grid dissipation (watts)	25
Capacitance ($\mu\mu$f) C_{fp}	6
C_{gf}	7
C_{pf}	0.4

acteristics are very uniform. They are given in Table II.

Constant-current curves are shown in Fig. 7. A typical curve of grid watts and primary grid current is shown in Fig. 8. A curve of current division between anode and grid for the same applied voltage on both electrodes is shown in Fig. 9. The average ratio of anode-to-grid current is 2.5 to 1.

A flow of water of 0.5 to 1 gallon per minute is necessary and sufficient to cool the tube in operation.

3. *Tests and Results*

Typical operating conditions for oscillator and amplifier are shown in Table III.

Most of the recent studies of this tube have been made at 600 megacycles with continuous-wave operation. At this frequency, it has been studied as an oscillator and as a neutralized amplifier in a grid-separation circuit. It has also been operated successfully as a doubler from 240

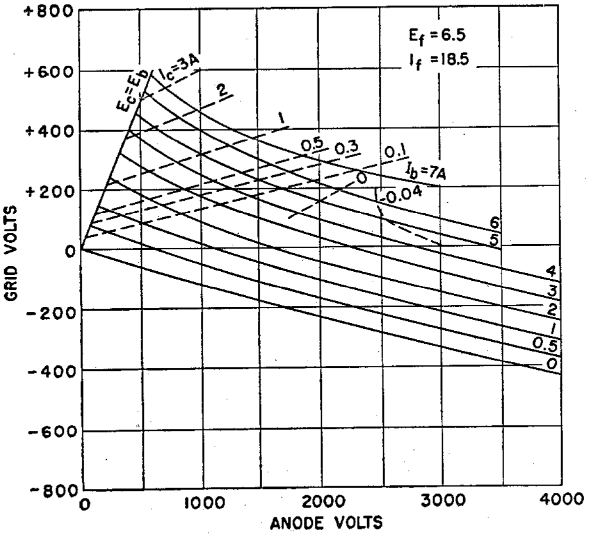

Fig. 7—Constant-current curves of the 6C22.

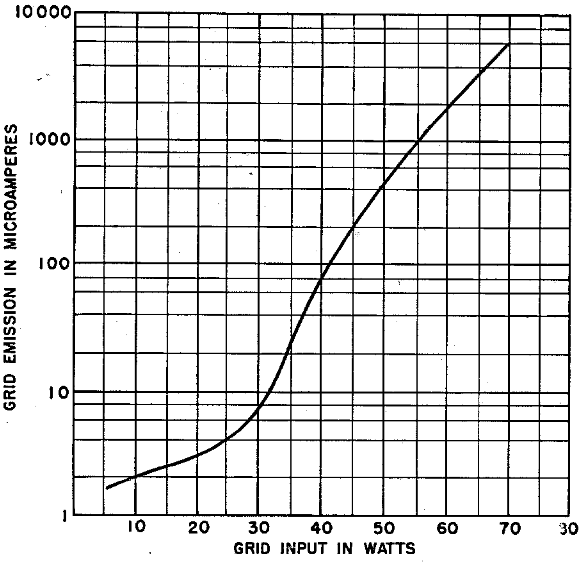

Fig. 8—Primary grid current plotted against grid input power.

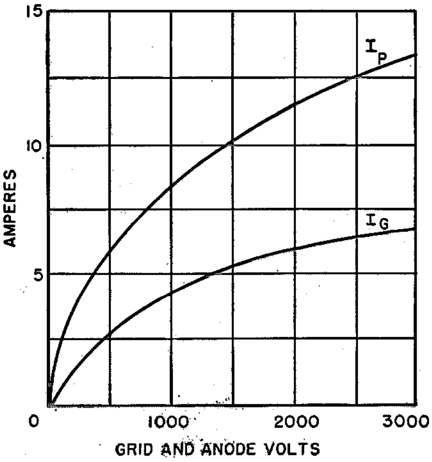

Fig. 9—Current division between anode and grid with the same voltage applied to both electrodes.

TABLE III

Typical Experimental Operation of 6C22

	Oscillator	Neutralized Inverted Amplifier
Frequency (megacycles)	600	600
Anode direct volts	1200	1600
Anode direct current (amperes)	0.6	0.65
Anode power input (watts)	—	1040
Grid direct current (amperes)	0.050	—
Power output (watts)	250	500
Driving power (watts)	—	190
Power gain	—	2.6
Anode dissipation (watts)	470	—
Efficiency (percent)	35	—

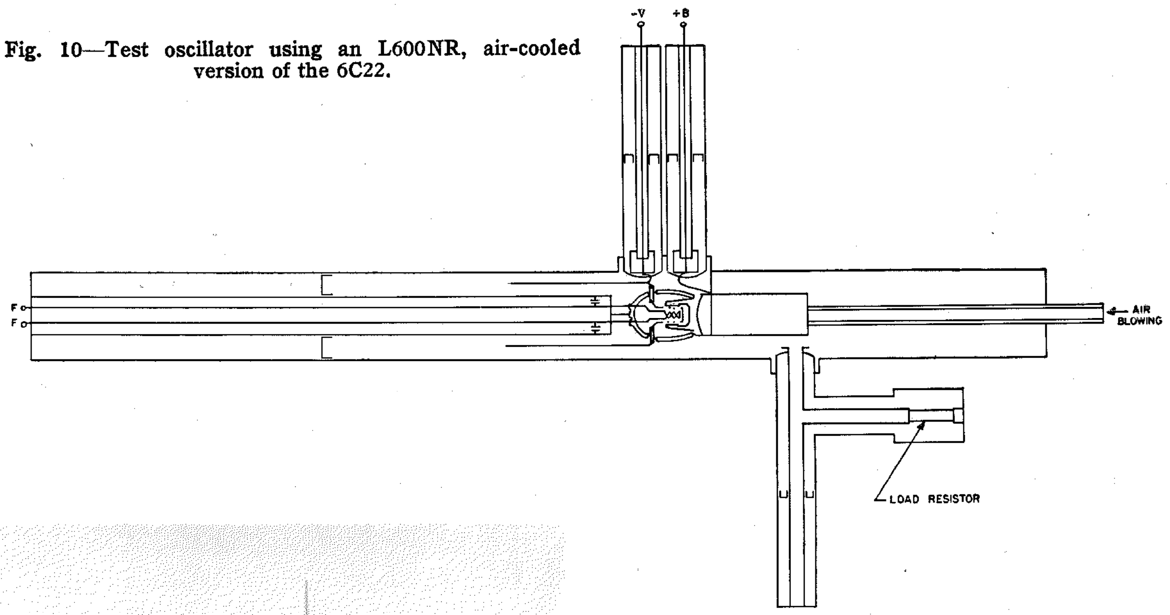

Fig. 10—Test oscillator using an L600NR, air-cooled version of the 6C22.

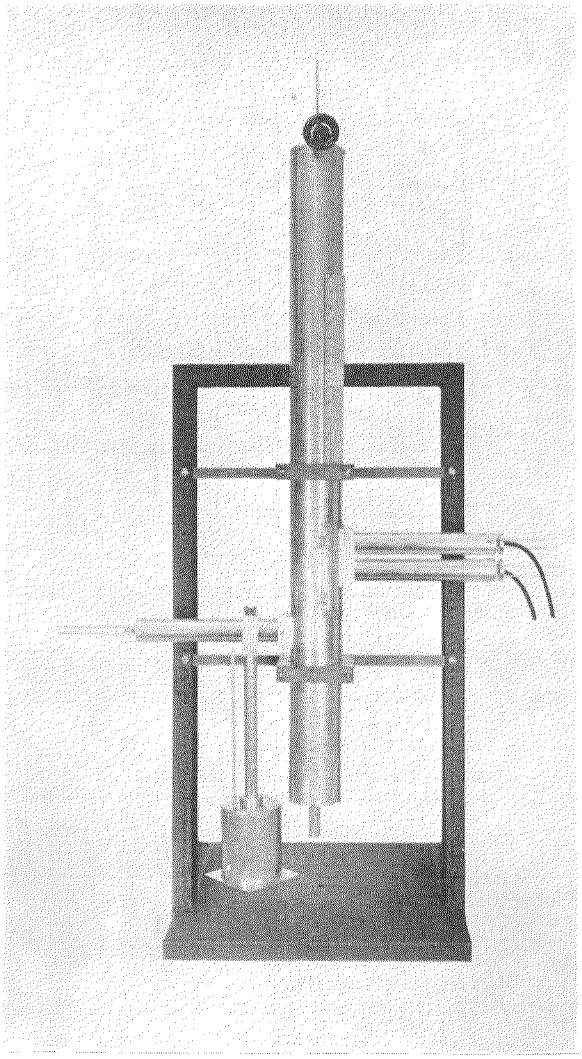

Fig. 11—Experimental setup of test oscillator with calorimeter load.

to 480 megacycles and as a tripler from 200 to 600 megacycles. The performance data under these conditions are given in Table IV.

In all cases, the circuits used are of the coaxial type to assure uniform current distribution and thereby reduce losses to a minimum. This is

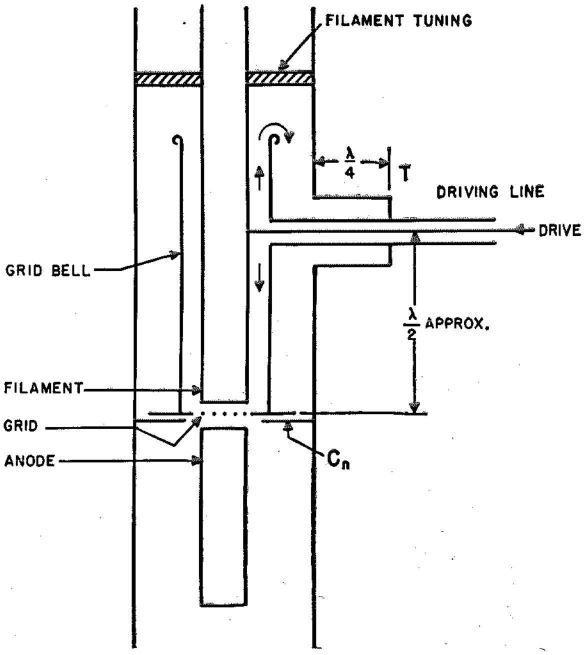

Fig. 12—Neutralization arrangement.

desirable in any case, but is particularly necessary to protect the grid seal from over-heating caused by high current concentration.

A diagrammatic view of a test oscillator using an air-cooled developmental version of the 6C22 known as the L600NR is shown in Fig. 10. A view of the experimental set-up with calorimeter load is shown in Fig. 11. Anode voltage and grid bias voltage are brought in through quarter-wave chokes. The cathode circuit is piston tuned, while the anode is adjusted for "half-wave-length" open-line operation. Radio-frequency output is coupled to the load by means of a capacitive pickup and matching section.

A hollow brass cylinder, connected at one end to the grid ring, open at the other end, and commonly termed a "grid bell," is adjustable in length, and serves to determine the amount and phase of feedback from output to input circuits.

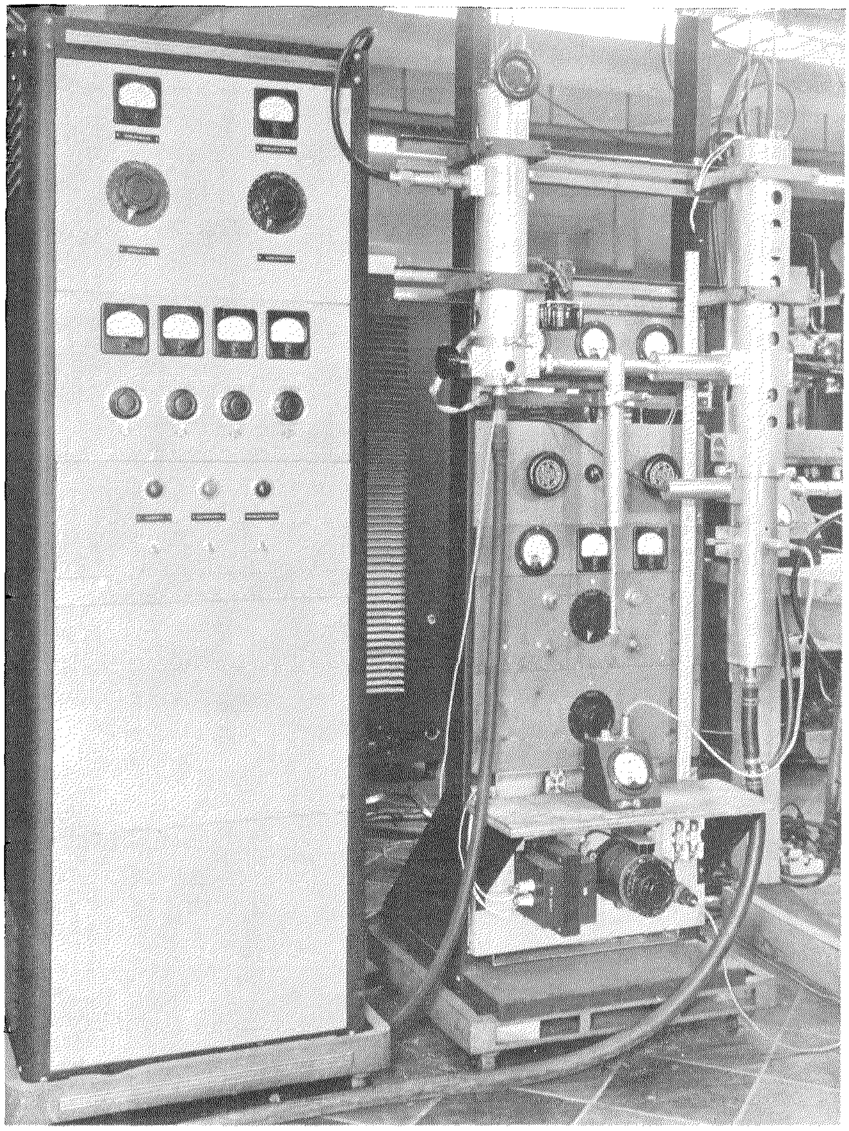

Fig. 13—Laboratory model of crystal-controlled transmitter delivering 500 watts at 600 megacycles. The exciter operates a 6C22 tripler at the left which drives the final 6C22 amplifier at the right.

Cooling air is brought to the anode through a dielectric pipe which extends into the inner region of the anode-line cylinder. In the case of the 6C22, which is liquid cooled, two dielectric pipes of small diameter are used.

TABLE IV

Typical Operating Conditions for 6C22

	Frequency Doubler	Frequency Doubler	Frequency Tripler
Input frequency (megacycles)	240	240	200
Output frequency (megacycles)	480	480	600
Anode direct volts	1500	1700	1100
Anode direct current (amperes)	0.540	0.640	0.275
Anode power input (watts)	810	1080	—
Grid direct volts	−430	−430	−400
Grid direct current (amperes)	0.025	0.025	0.020
Power output (watts)	285	360	100
Driving power (watts)	—	—	150
Efficiency (percent)	35	33	33

When operated as a neutralized grounded-grid amplifier, results are obtained as given in Table III. Neutralization may be considered to have a twofold purpose. The first purpose is to reduce regeneration to prevent oscillation or to eliminate feedback through the tube entirely. The second purpose is to reduce interaction between the amplifier output circuit and its driving circuit to a minimum.

At lower frequencies these conditions can be satisfied simultaneously by the adjustment of a single parameter because the output terminals of the driver and the input terminals to the active elements of the grid-cathode structure can be considered electrically identical. At ultra-high frequencies, where the impedance in the tube leads prevents access to the active tube electrodes, this simplification is no longer valid. Consequently, in general, two separate adjustments are required.

When these requirements, together with the desirability of coaxial-type structures, are taken into consideration, neutralization arrangements such as the example shown diagrammatically in Fig. 12 result. Here C_n is a comparatively large capacitance which permits a small amount of additional coupling between the input and output structures. Simultaneously, the length of the grid bell is adjusted so that the resulting feedback voltage is zero as determined by a null reading in a detector inserted in the driving line. In this manner, the two requirements for neutralization are satisfied.

The amplifier can deliver approximately 500 watts at 600 megacycles when driven either from a doubler or a tripler, or from another amplifier using the same type of tube. Fig. 13 shows a laboratory crystal-controlled transmitter delivering 500 watts at 600 megacycles. The cabinet on the left is an exciter with the following tube complement:

807 crystal oscillator and tripler (to 12.5 megacycles)
807 doubler (25 megacycles)
807 doubler (50 megacycles)
HK54 doubler (100 megacycles)
6C22 doubler (200 megacycles)

This exciter drives the 6C22 tripler shown on the rack at the left to deliver driving power to the 6C22 final amplifier (on rack at right) at 600 megacycles.

A neutralized water-cooled amplifier using a 6C22 has been grid-modulated satisfactorily in a television transmitter with video-frequency components up to 10 megacycles at a carrier frequency of approximately 500 megacycles. The synchronizing peak power output was 1 kilowatt.

Satisfactory life tests have been conducted on the 6C22 as a continuous-wave oscillator at 535 megacycles with 725 watts input, 35 percent efficiency, for 500 hours.

Vibration tests in both the horizontal and vertical mounting positions show no failures up to 11 g, with 120 watts of filament and 20 watts of grid power applied during the test.

4. *Acknowledgment*

Acknowledgment is made of the contributions to this development by P. G. Chevigny, who originated the design of the tubes, and to G. Lehmann for theoretical analysis of the tubes and circuits.

Standard Telephones and Cables Issues Second Edition of Reference Data for Radio Engineers

In 1942, Standard Telephones and Cables, Ltd., London, produced a small handbook, "Reference Data for Radio Engineers," catering primarily for the needs of British engineers. In 1943, Federal Telephone and Radio Corporation produced a companion volume under the same title, but primarily intended to meet American requirements. Both books were so well received that second, and considerably enlarged, editions were published in both countries in 1946.

The compilers of these two volumes, working independently, but within the common framework of the I.T.&T. System, have of course been able to take advantage of each other's work, and while each volume has an individuality in style and arrangement corresponding to its country of origin, both volumes also bear the System stamp in respect of the breadth and technical treatment of the fields with which they deal.

Triodes for 3- and 10-Kilowatt Frequency-Modulated Transmitters

By P. I. CORBELL, JR. and H. R. JACOBUS

Federal Telephone and Radio Corporation, Newark, New Jersey

TRANSMITTING TRIODES, 7C26 and 7C27, designed for low and medium power in the new frequency-modulation broadcasting band are described. They permit from 1 to 10 kilowatts of output power to be generated with a minimum number of triode amplifier stages. Mechanical and electrical design features are discussed.

• • •

Assignment of the frequency band from 88 to 108 megacycles per second by the Federal Communications Commission to frequency-modulation broadcasting resulted in a need for a completely new line of power transmitting tubes. The 7C26 and the 7C27 were designed for the output stages of 3- and 10-kilowatt transmitters, respectively.

The smaller tube, type 7C26, will give a power output of 1.5 kilowatts and is designed for use in a grounded-filament circuit, as neutralization at low power is not difficult. It is of conventional design with the concomitant advantages of low cost and ease of production.

At the next higher power level, neutralization becomes more difficult because of the higher voltages and capacitances encountered. Therefore, the design of the 7C27 was based on its use in grounded-grid circuits, which do not require neutralization. A pair of 7C27's will easily deliver

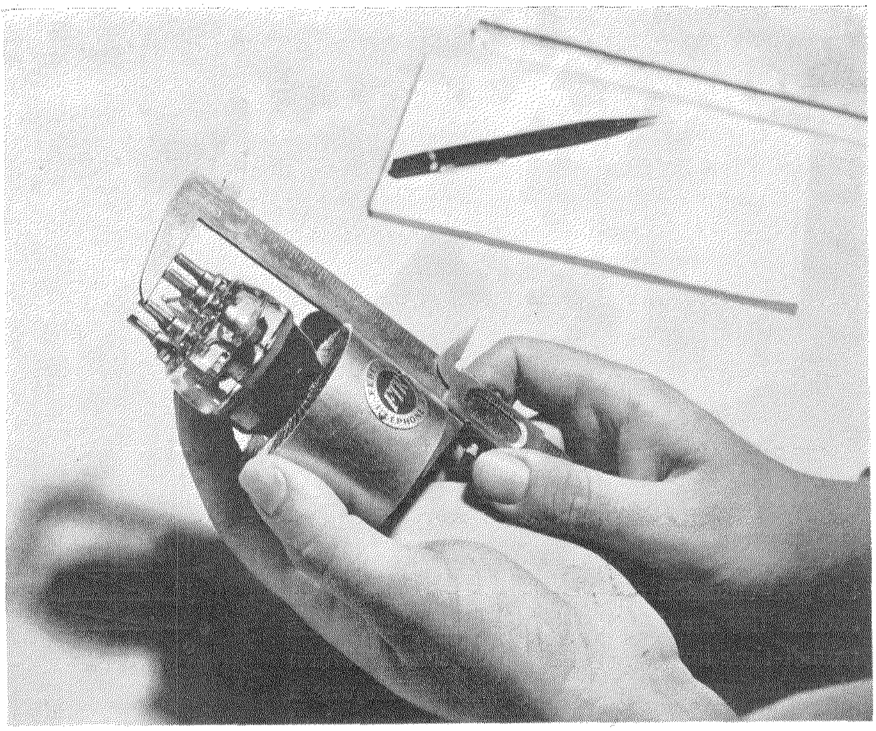

Fig. 1—7C26, air-cooled triode capable of giving 1.5 kilowatts of power at 150 megacycles. The overall length is 4$\frac{3}{8}$ inches, and the diameter is approximately 2$\frac{1}{2}$ inches.

10 kilowatts of output power when driven by a pair of 7C26's.

The operating frequencies impose certain design limitations and introduce problems not encountered in low-frequency work. At these frequencies, tuned circuits are resonant lines because the required small values of inductance and capacitance are impractical in lumped-constant form. The tube interelectrode capacitances that shunt these lines are an appreciable portion of the total circuit capacitances. Similarly, the inductance of the tube electrodes and leads contributes substantially to the total circuit inductance. It is evident that those distributed impedances affect over-all circuit performance and must be minimized to attain satisfactory operation.

The physical length of the two tubes has been reduced to the minimum, consistent with such other design considerations as voltage stress on the glass and heat dissipation. The grid-support structures are cylindrical or cone shaped, have low inductance, and provide good shielding between filament and anode.

Kovar-to-glass seals have been utilized wherever possible because they provide a strong mechanical structure at low production cost.

Forced-air-cooling has been provided by using an adequate and convenient radiator consisting of radial copper fins silver soldered between two copper shells.

At 100 megacycles, electron transit time must be considered. The time taken by an electron to traverse the grid-filament space may approach the period of one radio-frequency cycle if the grid-filament spacing is large. This effects a phase shift in the output current and voltage, and reduces efficiency. Some electrons, accelerated late in the cycle, may be returned to the filament with a resulting loss of radio-frequency output power. It has been found experimentally that tube operating efficiency will not be adversely affected by a total transit time of the order of one-tenth the period of the applied signal.

Close electrode spacings serve to aggravate the grid-emission problem, and it is desirable to choose the maximum spacing dictated by the transit-time limitation. Even with the maximum allowable spacing, primary grid emission was a very serious problem, and exhaustive tests had

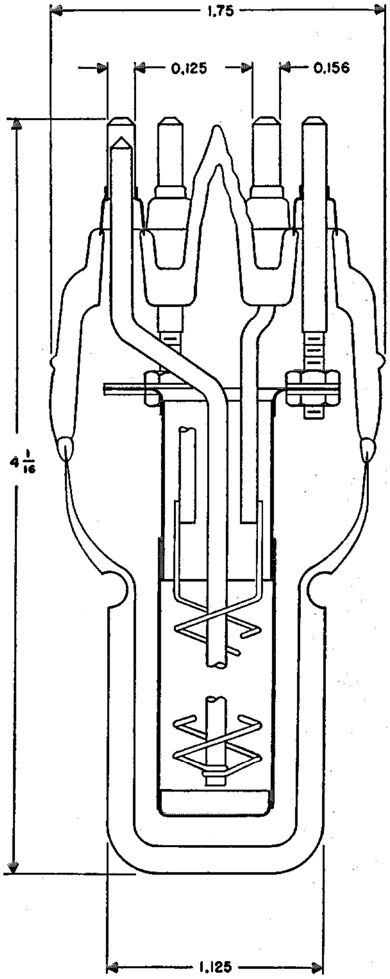

Fig. 2—Plan drawing of the 7C26. There are six base pins, three of which are connected by short leads to the grid structure. The other three are for the thoriated-tungsten filament and its center tap. The radiator is not shown.

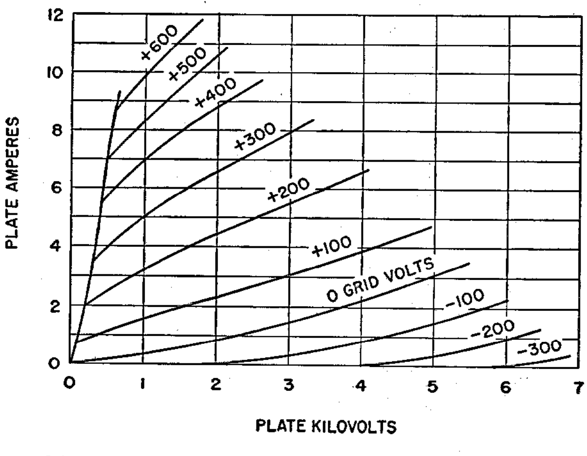

Fig. 3—Average plate characteristics of the 7C26.

to be made on various grid materials. After selecting a satisfactory material, it was necessary to develop suitable methods of processing with particular emphasis on minimizing grid contamination during exhaust. The resulting grid structures are practicable to build and possess electrical characteristics that represent a good compromise between extremely high perveance and negligible emission. In addition, these grids have been made extremely rigid so that uniformity of characteristics may be maintained.

Both tubes have thoriated-tungsten filaments which operate at a lower temperature and require less heating power than pure tungsten for a given thermionic emission. The general characteristics of the tubes are given in Table I, and Table II shows maximum ratings and typical operating conditions.

TABLE I

GENERAL CHARACTERISTICS OF 7C26 AND 7C27

General Characteristics	7C26	7C27
ELECTRICAL		
Filament	Thoriated Tungsten	Thoriated Tungsten
Volts	9.0	16.0
Amperes	28.0	28.5
Amplification Factor	19	30
Direct Interelectrode Capacitances (micromicrofarads)		
Grid-Plate	10.0	14.0
Grid-Filament	11.0	14.0
Plate-Filament	1.0	0.4
MECHANICAL		
Type of Cooling	Forced Air	Forced Air
Minimum Flow (cubic feet per minute)	75	175
Maximum Over-All Dimensions (inches)		
Length	4.375	8.031
Diameter	2.55	3.532
Mounting Position	Vertical, Anode Down	Vertical, Anode Down
Direction of Air Flow	Up	Up

TABLE II

MAXIMUM RATINGS AND TYPICAL OPERATION OF 7C26 AND 7C27

Class-C Radio-Frequency Power Amplifier and Oscillator without Amplitude Modulation

Characteristic	Typical Operation		Maximum Rating	
	7C26	7C27*	7C26	7C27
Plate, direct volts	3000	4000	3000	4000
Plate Current (amperes)	0.75	1.5	1.0	2.0
Grid, direct volts	−800	−600	−800	−1000
Grid Current (amperes)	0.100	0.200	0.125	0.200
Plate Input (watts)	2250	6000	3000	8000
Plate Dissipation (watts)	680	2350	1000	3000
Power Output (watts)	1570	5150	—	—
Maximum Frequency (megacycles)	108	108	150	110

* In push-pull grounded-grid amplifier with 3-kilowatt driver. All values for one tube.

1. Type 7C26

Type 7C26, shown in Fig. 1, is of conventional construction. Its short structure reduces inductance to a value where grid-ring construction is unwarranted in view of accompanying increased production costs. Instead, three leads are brought from the grid cone to terminal pins at the top of the envelope, permitting symmetry in external connections. This multiple connection, shown in Fig. 2, is essentially a low-inductance arrangement, and permits a balanced distribution of radio-frequency currents to the grid structure. The leads are threaded and connection to the grid cone is by means of nut-and-washer assemblies. This method allows accurate grid-to-filament alignment without demanding subassemblies of unusually close tolerances.

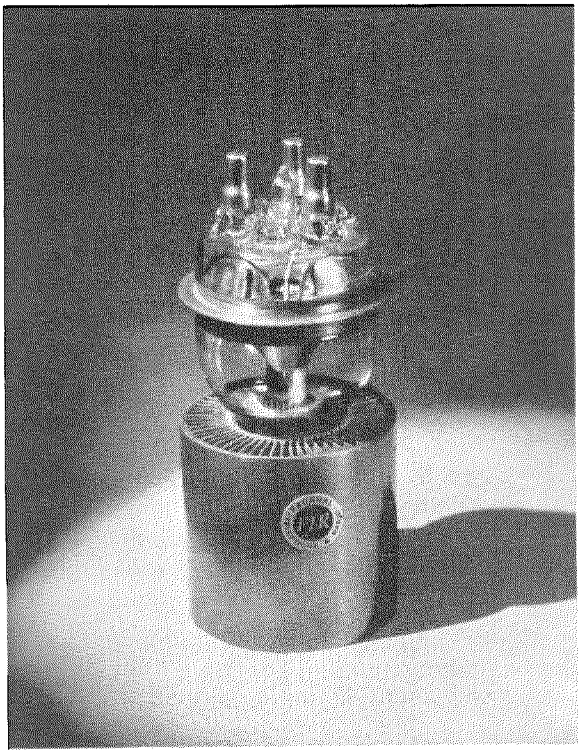

Fig. 4—7C27 air-cooled triode capable of giving 5 kilowatts of power at 110 megacycles.

The filament is a double helix with a center-tap supporting member; external connections are symmetrically located with respect to the grid terminals as may be seen in Fig. 2.

The location of these terminals at the top of the tube envelope permits use of a molded glass dish with filament and grid leads fixed in position by kovar sealing cups.

The anode is grooved, as may be noted in Fig. 2. The reduced wall thickness of this section offers resistance to heat flow toward the anode seal area. The main anode body below the groove may be operated during exhaust at a uniformly high temperature while the seal area, subjected to an air blast, remains cool. This permits very thorough anode degassing. An air flow of 75 cubic feet per minute through the radiator allows operation at full ratings. The plate characteristics are shown in Fig. 3.

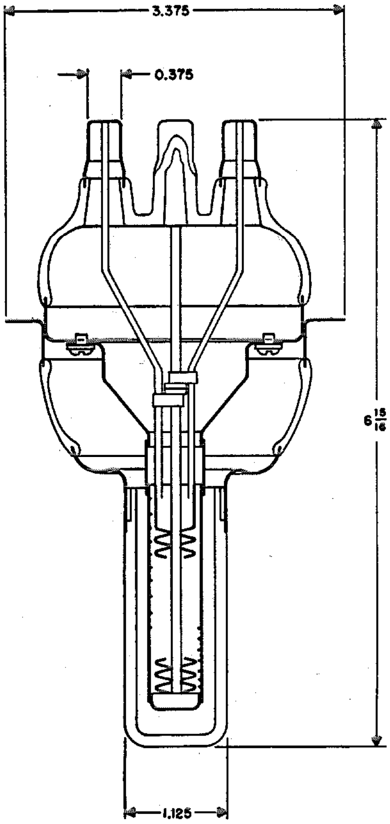

Fig. 5—Plan drawing of the 7C27. The radiator is not shown.

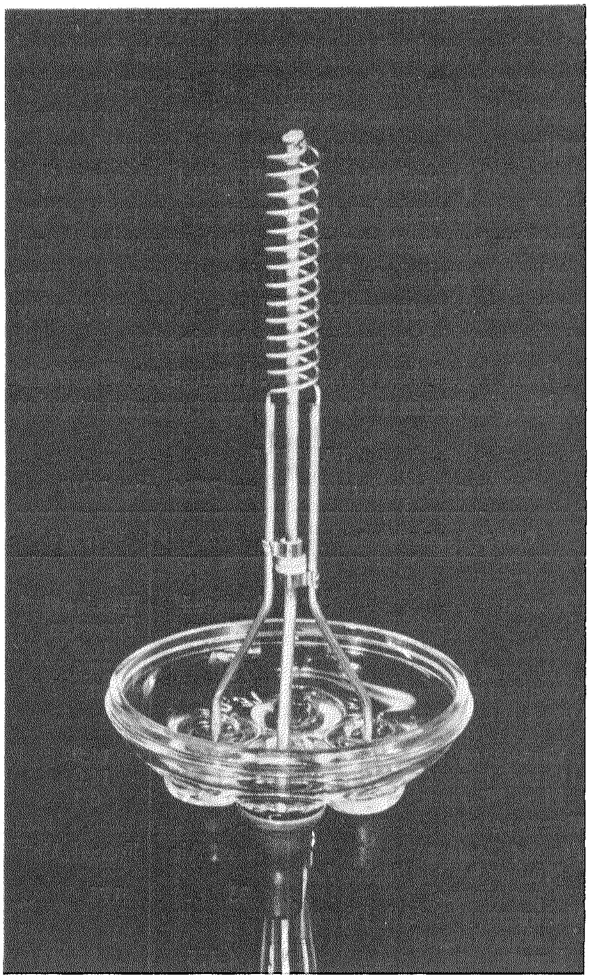

Fig. 6—Filament assembly of the 7C27. The filament leads are strapped to the heavy center-tap rod for rigidity. Small ceramic insulators are used under the straps.

These tubes have been operated in a push-pull oscillator circuit at 280 megacycles at approximately half ratings and have oscillated at 340 megacycles with lower outputs. At these frequencies, however, the transit-time effects, in addition to decreasing the power output, may seriously affect filament life because of electron back-bombardment.

2. *Type 7C27*

Type 7C27, a photograph of which is reproduced in Fig. 4, will provide 5 kilowatts of output power at 110 megacycles. As the tube design includes a grid ring and a conical grid support, good filament-plate shielding is obtained, and the tube is very satisfactory in grounded-grid circuits. As may be seen in the general assembly

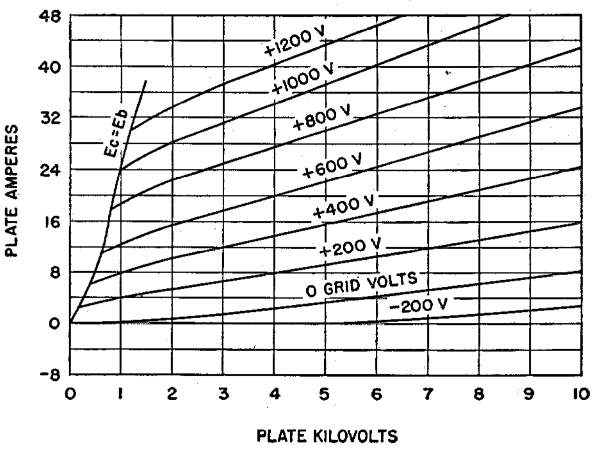

Fig. 7—Average plate characteristics of the 7C27.

drawing, Fig. 5, the tube has a slightly reentrant anode. This serves two purposes. It increases the length of the glass bulb between grid and anode without increasing the over-all tube length, and it eliminates sharp glass contours between these electrodes by providing a large diameter for the anode sealing edge. All glass-to-metal seals are of the kovar rounded-edge type. The grid support consists of a copper cone fused directly to a molybdenum band, using a special radio-frequency brazing technique. This combination of materials aids in simplifying the problem of degassing the grid structure and minimizes radio-frequency losses.

The filament-supporting structure, Fig. 6, is short and simple. Rigidity is improved by mechanically strapping the two leads to the center support rod below the spiral filament. A small ceramic spacer insulates the straps. The center tap provides a connection to the external circuits.

The plate characteristic curves are shown in Fig. 7.

3. Acknowledgment

Acknowledgment is herewith given to the Federal Telecommunication Laboratories for the initial electrical design. Appreciation is extended to Messrs. A. K. Wing, Jr.; W. Happe, Jr.; H. W. Baker; and J. Budig for assistance and suggestions during the development of these tubes.

Insertion Loss and Effective Phase Shift in Composite Filters at Cut-Off Frequencies

By VITOLD BELEVITCH

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THE USUAL METHODS of estimating the insertion loss of a nondissipative filter as a sum of several terms (reflection and interaction losses) fail at the theoretical cut-off frequencies because the reflection losses are infinite, whereas the interaction loss is negatively infinite. By direct calculation, however, very simple formulas for the total loss are obtained, at least in the most practical case of filters having equal or inverse image impedances at both ends. These formulas are also of practical value for dissipative filters, although the usual method could then be applied, because the supplementary dissipative loss is generally small compared to the loss in a nondissipative filter.

All the preceding considerations hold good for the calculation of effective phase shift at cut-off frequencies.

In the first part of this paper, formulas and notations are listed. In the second part, proofs are developed for the main cases.

• • •

1. *Low-Pass-Filter Formulas*

Let us first consider the case of a low-pass filter having equal image impedances and working between equal ohmic resistances at both ends. The image impedances may be of the constant-k or m-derived type, either mid-series or mid-shunt. The following usual notations are employed:

f_c = cut-off frequency.
$f_1, f_2, \ldots$ = frequencies of infinite image attenuation.
$m_1, m_2, \ldots$ = corresponding parameters of Zobel, defined by

$$m_i = \frac{(f_i^2 - f_c^2)^{\frac{1}{2}}}{f_i}. \tag{1}$$

$m_t = m$ corresponding to the terminal m-derived image impedances; $m_t = 1$ if image impedances are of the constant-k type.
μ = ratio of the image impedances at zero frequency to the terminating resistances, if the image impedances are of the mid-series type; inverse ratio, if the image-impedances are of the mid-shunt type.
A_c = insertion loss in decibels at cut-off frequency.
B_c = effective phase shift in radians at cut-off frequency.

We then have the formulas

$$A_c = 10 \log_{10} \left[1 + \frac{m_t^4}{\mu^2} \left(\frac{1}{m_1} + \frac{1}{m_2} + \ldots \right)^2 \right], \tag{2}$$

$$B_c = n\pi - \tan^{-1} \frac{m_t^2}{\mu} \left(\frac{1}{m_1} + \frac{1}{m_2} + \ldots \right), \tag{3}$$

where n is the total number of entire sections of which the filter is composed; the principal value of $\tan^{-1}$ between 0 and $\pi/2$ must be considered.

These formulas are easily applied to symmetrical low-pass filters having no coincident frequencies of infinite attenuation. The number of terms in the sum appearing in (2) and (3) is then equal to the total number of *entire sections* composing the filter. If there is any multiple frequency of infinite attenuation, it must be considered with its order: for instance, if a filter be composed of two entire sections with $m = 0.5$, and one entire section with $m = 1$, the sum $1/m_1 + 1/m_2 + 1/m_3$ is $2/0.5 + 1/1 = 5$.

In lattice filters, frequencies of infinite attenuation may be imaginary or may occur in conjugate complex pairs. It will be seen from the subsequent proof that (2) and (3) still hold in those cases.

Consider now filters having inverse image impedances at both ends. More precisely, if the terminating resistances and the image-impedances are denoted by R_1, R_2, Z_1, Z_2, the condition

$$Z_1Z_2=R_1R_2 \tag{4}$$

will be assumed, thus producing the equality of the reflection losses at both ends. As in the symmetrical case, both Z_1 and Z_2 are supposed to be either of the constant-*k* or of the *m*-derived type. Formulas (2) and (3) are still valid, but it must be observed that the filter now necessarily contains a half section of the constant-*k* type. This must appear as a term 1/2 in the sum $1/m_1+1/m_2+\ldots.$ and must also be taken into account in the value of n.

By frequency transformation, (2) and (3) are easily extended to high-pass filters and to band-pass filters having a symmetrical characteristic on a logarithmic frequency scale.

2. *Band-Pass Formulas*

For band-pass filters having different attenuation characteristics on either side of the transmitted band, new formulas must be developed, as the insertion loss and effective phase shift will be different at either cut-off frequency. Consider only the case of a symmetrical filter having a constant-*k* impedance at both ends and working between pure equal resistances.

The following notations apply:

f_{-c}, f_c = cut-off frequencies ($f_{-c}<f_c$).

$f_1, f_2, \ldots.$ = frequencies of infinite attenuation.

$m_1, m_2, \ldots.$ = corresponding parameters defined by[1]

$$m_i=\left(\frac{f_{-c}}{f_c}\times\frac{f_i^2-f_c^2}{f_i^2-f_{-c}^2}\right)^{\frac{1}{2}}. \tag{5}$$

$$\rho=(f_c/f_{-c})^{\frac{1}{2}}+(f_{-c}/f_c)^{\frac{1}{2}}. \tag{6}$$

μ = ratio of the image impedances at the frequency $(f_{-c}f_c)^{\frac{1}{2}}$ to the terminating resistances, if the image impedances are of the mid-series type; inverse ratio if they are of the mid-shunt type.

A_{-c}, A_c = insertion loss at frequencies f_{-c}, f_c.

B_{-c}, B_c = effective phase shift at frequencies f_{-c}, f_c.

The formulas are:

$$A_{-c}=10\log_{10}\left[1+\frac{1}{\mu^2\rho^2}(m_1+m_2+\ldots.)^2\right], \tag{7}$$

$$A_c=10\log_{10}\left[1+\frac{1}{\mu^2\rho^2}\left(\frac{1}{m_1}+\frac{1}{m_2}+\ldots.\right)^2\right], \tag{8}$$

$$B_{-c}=B'_{-c}+\tan^{-1}\frac{1}{\mu\rho}(m_1+m_2+\ldots.), \tag{9}$$

$$B_c=B'_{-c}-\tan^{-1}\frac{1}{\mu\rho}\left(\frac{1}{m_1}+\frac{1}{m_2}+\ldots.\right), \tag{10}$$

where B'_{-c} and B'_c are the image phase constants at the cut-off frequencies.

3. *Derivation of Formulas*

It is known that a symmetrical filter, having an image transfer constant θ and image impedances W, is equivalent to a lattice structure composed of impedances

$$Z_1=W\,\text{cotanh}\,\theta/2, \qquad Z_2=W\tanh\theta/2. \tag{11}$$

The insertion loss of a lattice network between resistances R is given by the formula

$$A=20\log_{10}\left|\frac{(Z_1+R)(Z_2+R)}{(Z_2-Z_1)R}\right|. \tag{12}$$

As both Z_1 and Z_2 are purely reactive, the absolute value can be calculated. Using the notation

$$q=\text{cotanh}\,\theta/2, \qquad w=W/R, \tag{13}$$

the insertion loss becomes

$$A=10\log_{10}\left[1+\left(\frac{jq}{w}\,\frac{1-w^2}{1-q^2}\right)^2\right], \tag{14}$$

a j factor having been artificially added to the terms enclosed in brackets to make jq/w a real quantity.

The transfer constant θ of a symmetrical low-pass filter may be considered to be the sum $\theta_1+\theta_2+\ldots.$ of the transfer constants of partial filters, each having a single attenuation peak. By (13),

$$e^\theta=\frac{q+1}{q-1},$$

---

[1] This is not the usual definition of m introduced by Zobel, but differs from it by a constant factor so chosen as to give inverse m for attenuation poles symmetrically located with respect to the midband frequency.

and we may write

$$e^{\theta} = e^{\theta_1} e^{\theta_2} \ldots., \text{ or } \frac{q+1}{q-1} = \frac{q_1+1}{q_1-1} \times \frac{q_2+1}{q_2-1} \times \ldots. \quad (15)$$

Each partial q function is obtained by multiplying the q_0 function of a constant-k section by the m parameter corresponding to the attenuation peak, i.e., by the formula

$$q_1 = m_i q_0 = m_i \frac{\Omega}{(\Omega^2-1)^{\frac{1}{2}}}, \quad (16)$$

where Ω stands for the ratio f/f_c. Solving (15), we have

$$q = \frac{M_n\, q_0^n + M_{n-2}\, q_0^{n-2} + \ldots.}{M_{n-1}\, q_0^{n-1} + M_{n-3}\, q_0^{n-3} + \ldots.}, \quad (17)$$

where the M's are the fundamental symmetrical functions of the m_i: $M_0 = 1$; $M_1 = m_1 + m_2 + \ldots.$; $M_2 = m_1 m_2 + m_1 m_3 + m_2 m_3 + \ldots.$; $M_n = m_1 m_2 \ldots. m_n$.

As the insertion losses of inverse networks are equal, the m-derived image impedance may be assumed to be of the mid-series type.

$$w = \frac{W}{R} = \mu \frac{(1-\Omega^2)^{\frac{1}{2}}}{1-(1-m_i^2)\Omega^2}. \quad (18)$$

At the cut-off frequency, $\Omega = 1$, all quantities become zero or infinite because of the factor $(1-\Omega^2)^{\frac{1}{2}}$. The principal value of q is

$$M_n / M_{n-1} (\Omega^2 - 1)^{-\frac{1}{2}},$$

and that of w is $j\mu m_t^{-2}(\Omega^2-1)^{\frac{1}{2}}$.

$$\frac{M_{n-1}}{M_n} = \frac{1}{m_1} + \frac{1}{m_2} + \ldots. \quad (19)$$

By replacing q and w in (14) by these values, (2) is obtained.

The effective phase shift of a lattice structure is given by the formula

$$B = \arg \frac{(R+Z_1)(R+Z_2)}{R(Z_2-Z_1)}, \quad (20)$$

which may be transformed into

$$\tan B = \frac{1+w^2}{jw} \times \frac{q}{1+q^2} = \frac{1+w^2}{2w} \tan B', \quad (21)$$

where B' stands for the imaginary part of θ, i.e., the image phase constant. Formula (3) is then easily derived but it contains implicitly an arbitrary term $k\pi$. The exact value can only be determined by a continuous frequency variation starting from the zero frequency, where both B and B' are known to be zero. As B' is monotonically increasing and the factor involving w in (21) is always larger than 1, the following inequalities hold in the transmitted band:

$$\left.\begin{array}{ll} B' + \frac{\pi}{2} > B > B', & \text{if} \quad k\pi < B' < (k+\frac{1}{2})\pi, \\ B' > B > B' - \frac{\pi}{2}, & \text{if} \quad (k-\frac{1}{2})\pi < B' < k\pi, \end{array}\right\} \quad (22)$$

k standing for any integer. The determination of B_c given in (3) is deduced from the preceding inequalities taking into account the known value $n\pi$ of B' at cut off.

A similar derivation is possible in the case of filters having inverse image impedances, if (4) is satisfied. It may be shown that (14) still holds if q is replaced by

$$q' = \operatorname{cotanh}\left(\frac{\theta}{2} \pm i\frac{\pi}{4}\right). \quad (23)$$

For band-pass filters, similar results are obtained by using the decomposition (15) with the partial functions

$$q_i = m_i \left(\frac{f_c}{f_{-c}} \times \frac{f^2 - f_{-c}^2}{f^2 - f_c^2}\right)^{\frac{1}{2}}, \quad (24)$$

where m_i is given by (5). The mid-series constant-k image impedance is

$$w = \frac{\mu[(f^2 - f_{-c}^2)(f_c^2 - f^2)]^{\frac{1}{2}}}{f(f_c - f_{-c})}. \quad (25)$$

To overcome the indetermination arising in the phase expressions, they have been written as differences from the image phase constants.

The general validity of the formulas, even for filters with complex m's, proceeds from the decomposition formula (15), which was first proved by H. W. Bode[2] for band-pass filters and generalized by W. Cauer[3] who introduced the notations q and q' used in this paper.

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[2] H. W. Bode, "A General Theory of Electric Wave Filters," *Journal of Mathematics and Physics*, v. XIII, pp. 275–362; November, 1934.

[3] W. Cauer, "Theorie der linearen Wechselstromschaltungen," Akademische Verlagsgesellschaft Becker & Erler, Leipzig, 1941; p. 244.

Portable Direct-Reading Traffic Recorder

By AUGUST F. JONES

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ESTIMATES of the amount of equipment required for a telephone central office are based on the average number of calls and their duration for the busiest hour in the day. The portable direct-reading traffic recorder described in this paper will indicate the average occupancy of each of four groups of 50 traffic paths from which the traffic in 100- or 120-second calls is directly evident.

1. General Concept of Telephone Traffic

The amount of switching equipment required in a telephone exchange depends not only on the number of subscribers' lines served but also on the amount of traffic passing through the exchange in a given time. The traffic varies during the day and from day to day, but estimates of the amount of equipment needed are usually based on the "busy-hour traffic," that is, the average load during the busiest hour of a normal day. The number of calls that can be handled by a group of circuit paths during the busy hour is affected by the length of time each call occupies a path, or the "holding time." Thus traffic intensity involves both "calling rate" and "holding time."

Traffic is usually measured by an equivalent number of calls of a fixed holding time, two popular standards being the 100-second call and the 120-second call. The maximum capacity of a group of 10 paths during one hour is 360 100-second calls, or 36 100-second calls for each path. This assumes, however, that calls arrive at precisely timed intervals, and there is no time during which any path is idle. This is a false assumption, as indicated by the prediction of the formulas of probability that a group of 10 paths can handle 149 100-second calls occurring at random intervals during an hour (with a loss probability of 0.01).

2. Measurement of Telephone Traffic

As the number of paths to be provided, and hence the cost of the exchange equipment, depend on the "busy-hour traffic," it is of first importance to be able to measure traffic accurately and conveniently. A common method of measuring the number of calls is by use of "message registers," or meters, one connected to each path in such a way as to be operated each time the path is seized. The total number of calls during the busy hour is obtained by totaling the meter readings, and the average holding time is estimated by making observations with a stop watch, or by other means.

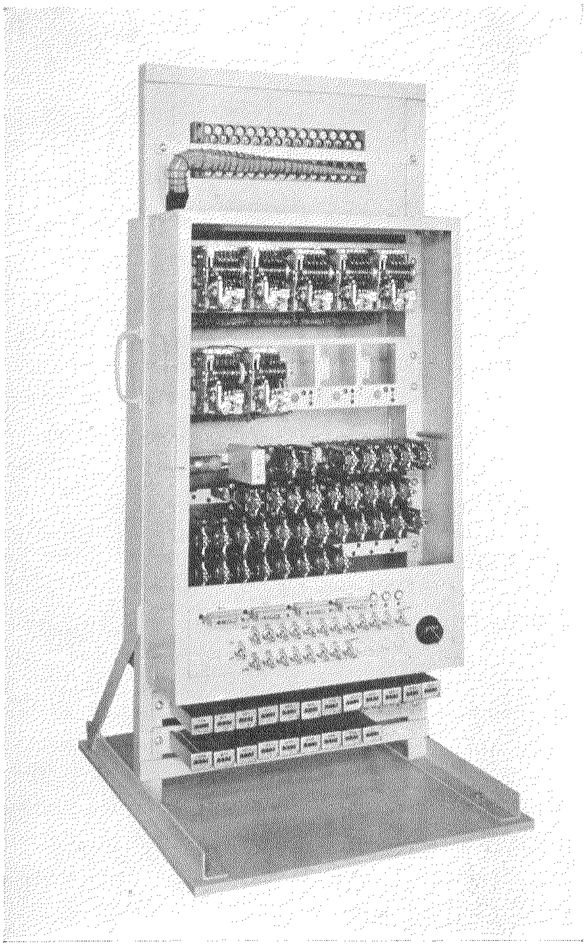
The equipment to be described in this article is designed to avoid any calculation or estimation of holding time and to give a direct reading of the traffic carried by a group or groups of circuit paths in 100-second (or 120-second) calls. It is based on the fact that the total traffic carried by a group of circuit paths during the busy hour is directly related to the average of the number of paths in use at any instant, or more specifically that the average occupancy of a group of paths, multiplied by 36, will give the number of equivalent 100-second calls carried by that group during the hour to which the average applies. For example, if during the busy hour an average of 5 of a group of 10 circuit paths is simultaneously engaged, the equivalent traffic carried by that group during the busy hour is 5×36=180 100-second calls.

3. Performance of the Traffic Recorder

The traffic recorder illustrated can measure the average occupancy of several groups of circuit paths, and give a direct reading in 100-second calls of the traffic handled by each group. Instead of making periodic checks during the busy hour of the number of paths of a group simultaneously occupied, it accomplishes the same result by going through rapidly repeated cycles of testing each path of a group for occupancy, one at a time, adding the total number of occupancies shown by these tests, and dividing by the number of test cycles performed to obtain the average occupancy of the group. This average occupancy is multiplied by 36 to give the traffic in 100-second calls. The machine accomplishes this computation by making exactly 360 test cycles. Each time a path is found busy, a message register assigned to the group of paths under observation is operated, and its complete reading shows the total number of occupancies encountered. Division of this number by 360 and multiplication by 36 are indicated by neglecting the last digit of the message register reading.

The recorder can measure directly the traffic on groups of as many as 50 paths. The tests for occupancy are made through five ratchet-driven selector switches connected in cascade. The message register is connected through the brush of the first switch successively to each of 10 contacts in the arc of the switch, then through the 11th contact to the brush of the 2nd switch, which steps successively over 10 contacts in its arc and then stops on its 11th contact to extend the connection to the brush of the 3rd switch, etc. The stepping is accomplished by means of periodic impulses generated in the recorder. The effect of the five 11-point switches operating in cascade is that of one large switch with an arc of 50 contacts, each contact connected to a separate circuit path. The five separate switches, however, give flexibility and permit traffic measurements of five separate groups of 10 paths each by the use of a separate message register connected to the brush of each switch.

Each of the above-mentioned switches has, besides control levels, 4 levels or rows of terminals to which can be connected circuit paths on which traffic is to be measured. Each level has its own brush and associated message register. Thus, in addition to the group of 50 paths observed as described above, 3 additional groups of 50, or 15 groups of 10, or any intermediate combination, can be observed. Traffic on a group of more than 50 paths can be observed by dividing the group into subgroups of 50 paths or less and adding together the readings of the subgroups. The number of paths in the groups being observed need not be in multiples of 10. If some of the arc contacts of the selector switches are not connected to circuit paths, tests in these positions will of course never cause an operation of the message register, and the traffic recorded will be that on the paths connected through terminals of the recorder to the selector arcs.



The impulse generator is calibrated to give an impulse every 200 milliseconds, and therefore effect a complete cycle of 50 steps every 10 seconds, or the series of 360 cycles in 1 hour. The impulse frequency does not affect the accuracy of the traffic-recorder readings, however, because these are based on average occupancy during the observation period. The only effect of changing the impulse frequency is to change the length of time over which the results of the 360 test cycles are averaged.

4. *Construction*

The illustration shows a front view of the portable traffic recorder. It is 38 inches high, weighs about 75 pounds, and is equipped with handles for convenience of transportation. At the top is the terminal strip to which connections are made from the paths to be observed. The capacity of the recorder is 200 paths, in groups not exceeding 50 each. The required connections of the message registers within the recorder are also made on this terminal strip. The 20 message registers provide for the maximum number of groups that can be observed at one time, i.e., one for each of the four brushes of each switch.

The 5 selector switches used for the occupancy tests are shown at the top of the equipment just under the terminal strip. Below these are two additional selector switches arranged to count the cycles and to stop the test after 360 cycles of 50 steps each. By changing a strap, the test can be stopped after 300 cycles and the readings will then be in 120-second calls. The 21st message register is used to record the number of cycles completed.

A relay and a toggle switch are associated with each message register. The relay controls the operating circuit of the message register, and the switch connects positive or negative potential to the winding of the relay, depending on which potential appears on the particular circuit path being observed to indicate a busy condition, or occupancy.

The impulsing circuit consists of relays, capacitors, and a variable resistor. The time constant of the circuit, and hence the frequency of impulsing, can be adjusted by the variable resistor, the knob of which is shown to the right of the toggle switches.

Operation of the starting switch causes all selector switches to return to normal position if not already there, and starts the impulsing circuit. The switches step automatically until the required number of cycles has been completed, after which the recorder shuts itself off.

5. *Acknowledgment*

This instrument was originally conceived by Mr. C. F. J. Boehlen, then a member of the Federal Telephone and Radio Corporation engineering staff. The recorder was designed and built under his direction and is now in regular use in an important telephone operating company in the U.S.A.

Comité Consultatif International Téléphonique

By P. E. ERIKSON

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THE Comité Consultatif International Téléphonique resumed its activities by holding a plenary meeting in London on October 29 and 30, 1945, at which 24 countries were represented. It was preceded by a meeting of technical and operating experts from various telephone administrations and private operating companies.

One of the most important problems concerned the restoration of international telephone service. The general consensus was that it should be done as quickly as possible so as to speed the solution of the problem of feeding the European population and aid in the reconstruction of destroyed property and equipment.

The plenary meeting unanimously recommended that: (A) a certain number of circuits, indispensable for the resumption of European service, should be established at once; (B) the use of radio circuits for overland traffic should be confined to cases where it is manifestly impossible to use wires; and (C) the latent, but heavy, demand for future facilities should be met by establishing a general program of international circuits designed to insure a rapid service of good quality.

The last-mentioned objective, it was pointed out, should be attained by the most modern methods developed for transmission and switching techniques. A special commission was appointed to study this problem.

The commission, known as the Mixed Commission on the European Toll Switching Plan, is composed of operating and technical experts from the various European telephone administrations. It held four meetings and submitted its report to the plenary meeting in Montreux, Switzerland, on October 26 to 30, 1946. The approved plan is to use high-speed multichannel carrier telephone circuits which, as far as possible, are to be carried in cables and laid so as to form a traffic ring around Germany. From the ring there will be traffic spurs radiating outward and inward to toll switching centers not located on the ring. Manual switching is visualized at the start, but rapid operating methods will be applied, one feature of which will be the use of a single toll ticket made out in the country where the call originates. It is also contemplated that it will be possible to route traffic over the ring in both directions in the event of a breakdown in any part of the ring.

The Mixed Commission prepared extensive intercountry circuit tables compiled from data supplied by the various administrations as their best estimates of the traffic to be handled in 1952, three years after the ring is expected to be put in service. On the basis of these data, there will be a considerable number of circuits in the ring. For example, it is estimated that, in 1952, there will be 300 circuits passing through Belgium and France.

The eight regular commissions of Rapporteurs and their subcommissions have resumed the study of questions which were pending since 1939, and have undertaken consideration of new questions that have arisen in connection with the European toll switching plan. It may, therefore, be said that the Comité Consultatif Internation Téléphonique is now operating on a pre-war basis with the same organization as before.

Investigation of Errors in Spaced-Collector Direction-Finder Systems

By T. H. CLARK

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PRINCIPAL sources of errors in design and installation of direction-finder equipment using spaced vertical collector systems are considered. Errors attributable to the site are enumerated. The calculable errors arising from the design of one directional pair are investigated mathematically for particular arrays. General conclusions are applicable as specific design principles.

The errors from combining two orthogonally placed directional pairs are investigated generally and quantitatively as are errors resulting from the use of long cables connecting the antennas to the goniometer. Errors stemming from mechanical misalignment are investigated.

• • •

The purpose of these studies is to determine the principal sources of error in design and installation of direction-finding equipment using spaced vertical collector systems.

The collector system for direction finders of this type usually consists of five vertical antennas placed at the corners and the center of a square. The antennas at diagonally opposed corners are connected so that their outputs are opposed. This results in a figure-of-eight reception diagram for this pair. Such a pair of antennas shown in Fig. 1 will be called a *directional pair.* Each directional pair is a receiving system in itself; however, its directional properties are not useful for direction finding unless a second directional pair is similarly placed on the other diagonal of the same square. The two directional pairs are to be considered as two separate receiving systems. The fifth antenna at the center of the square is a third independent receiving system, having omnidirectional characteristics, and is used for the determination of sense. When it is combined with either or both of the directional pairs, a cardioid diagram of reception plotted against direction will result.

In the examples to be cited below, a direction finder designed for the nominal frequency range of 1.5 to 30 megacycles per second is used. The complete direction finder consists of four arrays, each designed for maximum sensitivity and lowest errors in its particular frequency range. This design proved to be very conservative, and an actual frequency coverage much greater than that originally specified was secured. In Table I, the nominal and the extended frequency ranges for each array are shown.

The sources of errors may be divided conveniently into several groups:

A. Site
B. One directional pair
C. Combination of two directional pairs
D. Goniometer.

These will be investigated mathematically as well as practically, and general conclusions will be drawn as applicable.

1. Site Errors

The errors of the site may be classified as the results of three general effects:

A. Nonhomogeneous ground conductivity
B. Reflections from nearby conductors
C. Absorption obstacles in the path of the waves.

The above sources of errors have been investigated to a certain extent by various writers,

TABLE I

DIRECTION-FINDER SYSTEM USING 4 ARRAYS

Array	Frequency Range in Megacycles	
	Nominal	Extended
1	1.5– 3.3	1.25– 5
2	3.3– 7.5	2.5 –10
3	7.5–15	5 –20
4	15 –30	10 –40

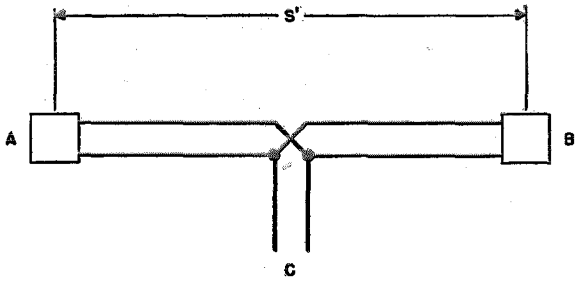

Fig. 1—A two-element array or directional pair equally applicable, from the standpoint of field pattern, to the transmitting or receiving cases.

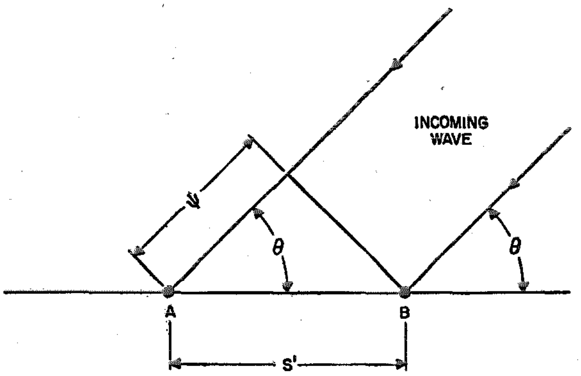

Fig. 2—Factors involved in calculating the receiving pattern of the two-element array (Fig. 1). θ = angle of reception to plane of array. S' = spacing angle = $360\ S/\lambda$, where S = actual spacing. $\psi = S' \cos\theta$.

as well as by the direction-finder group in these Laboratories.[1] Further investigation is very desirable since many of the problems are not yet clearly understood. The investigation of site errors is a complete study in itself, and will not be considered further in this paper as it is largely beyond the control of the equipment designer.

2. *One Directional Pair*

The errors which can be postulated in the design and construction of one directional pair will be investigated by a method which has been used previously to determine the field-strength diagrams of two-element transmitting arrays.[2]

Only those errors which result when the wave is arriving in the horizontal plane, i.e., when the vertical angle is zero, will be considered. The errors resulting from vertical angles of incidence other than zero are complicated by the horizontal-vertical reception ratio of the pair, and lead into another aspect of the subject which deserves special detailed treatment. For this study, therefore, consideration will be given only to the horizontally propagated field which the designer and installation crew can create for test purposes with a local source of emission such as a target transmitter placed on the ground at a distance of several wavelengths from the directional pair or array.

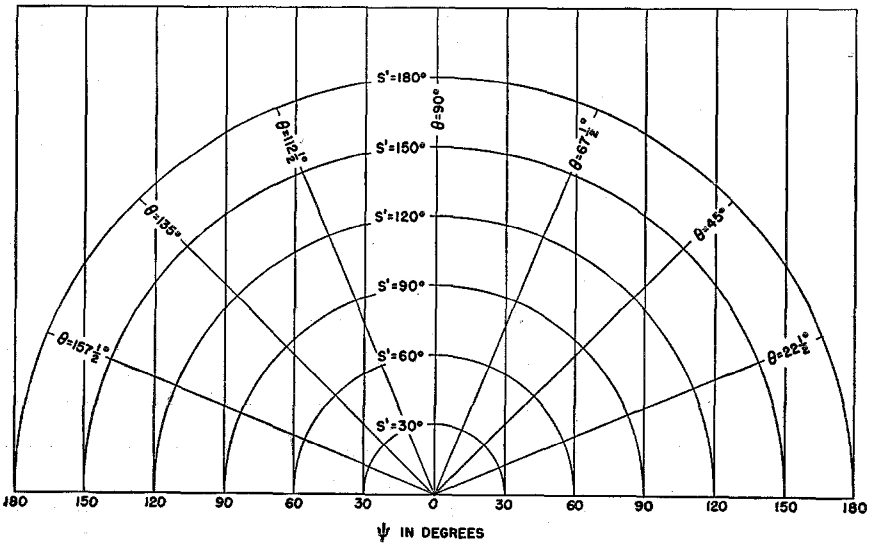

Fig. 3—Graph showing the relationship of ψ, S', and θ.

2.1 Directivity of Two-Element Array

The relative field strength at a given angle θ (Fig. 2) from the plane of such a two-element *transmitting* array is given by:

$$E_\theta = 1 + k \times (\cos\alpha + j \sin\alpha). \quad (1)$$

In other words, the relative field strength at any angle θ is equal to 1 plus a constant k at an angle α. This formula serves equally well for reception if the symbols are properly redefined.

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[1] H. Busignies, "Evaluation of Night Errors in Aircraft Direction Finding, 150–1500 Kilocycles," *Electrical Communication,* v. 23, pp. 42–62; March, 1946.

[2] E. A. Laporte, "Directional Antenna Design," *Electronics,* v. 9, pp. 22–25, 48; April, 1936.

Let k equal the ratio of the effective heights of the two elements. (To be discussed in more detail later.)

Let angle α be determined by the combination of the relative phases of the outputs from the two receiving elements at the point of cross connection; defined mathematically as

$$\alpha=\psi+\phi,$$

where ϕ is equal to the relative electrical phase from the two elements, A and B, at the terminal point C of the connecting transmission line (Fig. 1) when a wave arrives simultaneously on the two elements, i.e., when $\theta=90$ degrees.

Angle ψ is determined by the angle of arrival of the incoming wave, and is equal to $S'\cos\theta$, where θ, as stated above, is the angle of arrival of the incoming wave, and S' is the phase difference in electrical degrees between the voltages induced in the two elements for a wave arriving in the plane of the elements.

Distance S' in electrical degrees is equal to $360\ S/\lambda$. Here, S is the element spacing, and λ is the wavelength, both measured in the same units.

In the determination of the response pattern (or for the solution of (1) above) for a two-element antenna array, one may make use of short-cut graphical methods to determine the angle and resultant response E_θ.

2.1.1 Determination of ψ and α

In Fig. 3 we have plotted about a central point a number of circles having radii equal to S' in degrees. From this same center are laid off a number of radii whose angles, with respect to the horizontal axis, correspond to θ the angle of arrival of the waves. To determine ψ for any electrical separation S' and any angle of arrival θ, it is only necessary to determine the intersection of the circle corresponding to S' with the radius θ. The horizontal component of this point of intersection is equal to ψ in degrees. This will be evident from the mathematical treatment above. It is now necessary only to add the phase angle ϕ to the previously determined ψ to obtain α.

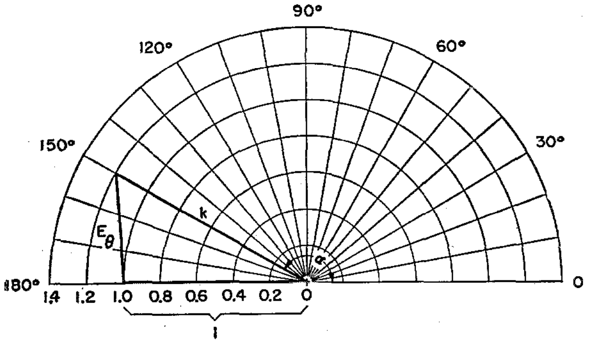

Fig. 4—Graphical solution of E_θ, where α and k are known.

2.1.2 Determination of E_θ

The magnitude of E_θ can be determined graphically by the means shown in Fig. 4. On polar-coordinate paper, a distance of one unit at an angle of 180 degrees has been laid off from the center. This represents the unit 1 in (1).

Now, lay off from the center the distance k at the angle α. The length of the line joining the two terminal points is E_θ.

The cases of most interest in direction finding occur when ϕ is very close to 180 degrees, and when k is equal to 1. A plot of E_θ for $k=1$ as a function of α in degrees is shown in Fig. 5. In

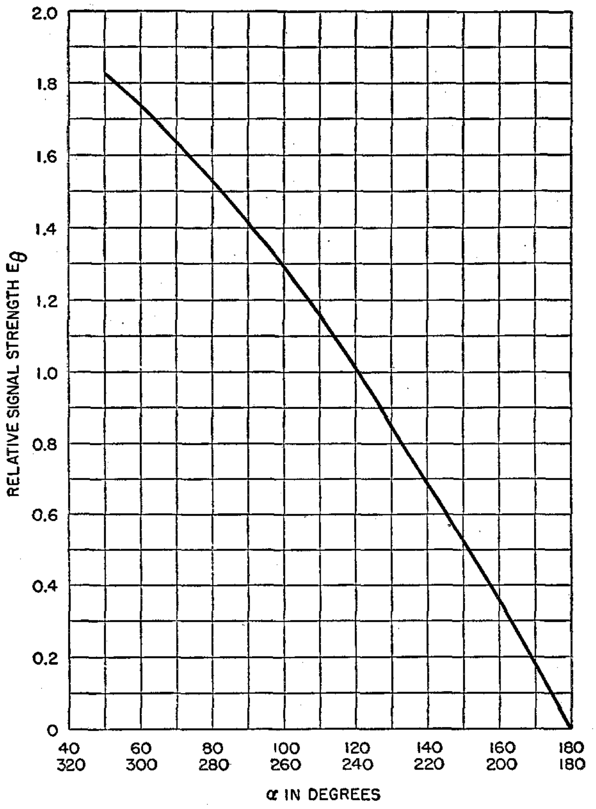

Fig. 5—Variation of E_θ with α for $k=1$.

Fig. 6 are plotted several field patterns which will be obtained for the conditions shown. Note that when ϕ differs by only a few degrees from 180 degrees, the nulls are quite visibly displaced. This case will be treated further below.

2.2 Vector Representation

In Fig. 7, vector addition has been performed at the crossover point. This point is represented by C in Fig. 1. The following discussion refers only to one directional pair.

Referring to Fig. 1, let the output from antennas A and B be E_1 and E_2, respectively. When a wave arrives in the null direction, E_1 and E_2 are in phase and, being connected in opposition, the resultant E_R is zero, if E_1 is equal to E_2 (i.e., $k=1$). This is shown vectorially in Fig. *7A*.

In Fig. *7B* we have shown the resultant E_R for an angle of wave arrival θ which causes E_1 to lead E_2. Note that E_R is at right angles to the positions occupied by E_1 and E_2 in Fig. *7A*. This condition is obtained when k is equal to 1 and θ is neither 90 nor 270 degrees.

At *7C* is shown a condition similar to that at *7B* with the exception that E_1 does not equal E_2. The resultant E_R is no longer at right angles to the previous positions of E_1 and E_2 as shown in *7A*.

At *7D* is shown the resultant of the same conditions as *7C*, except that the transmitter is now located at a position where $\theta = 90$ degrees. As E_1 is larger than E_2, E_R is now in the direction of E_1, and therefore its phase has changed by 90 degrees from the phase shown in *7B*. Note that even though the transmitter is in the desired null direction, there is still a resultant voltage. This is the common condition of blurred nulls.

In *7E*, we have shown the condition resulting from equal E_1 and E_2 voltages, but when ϕ does

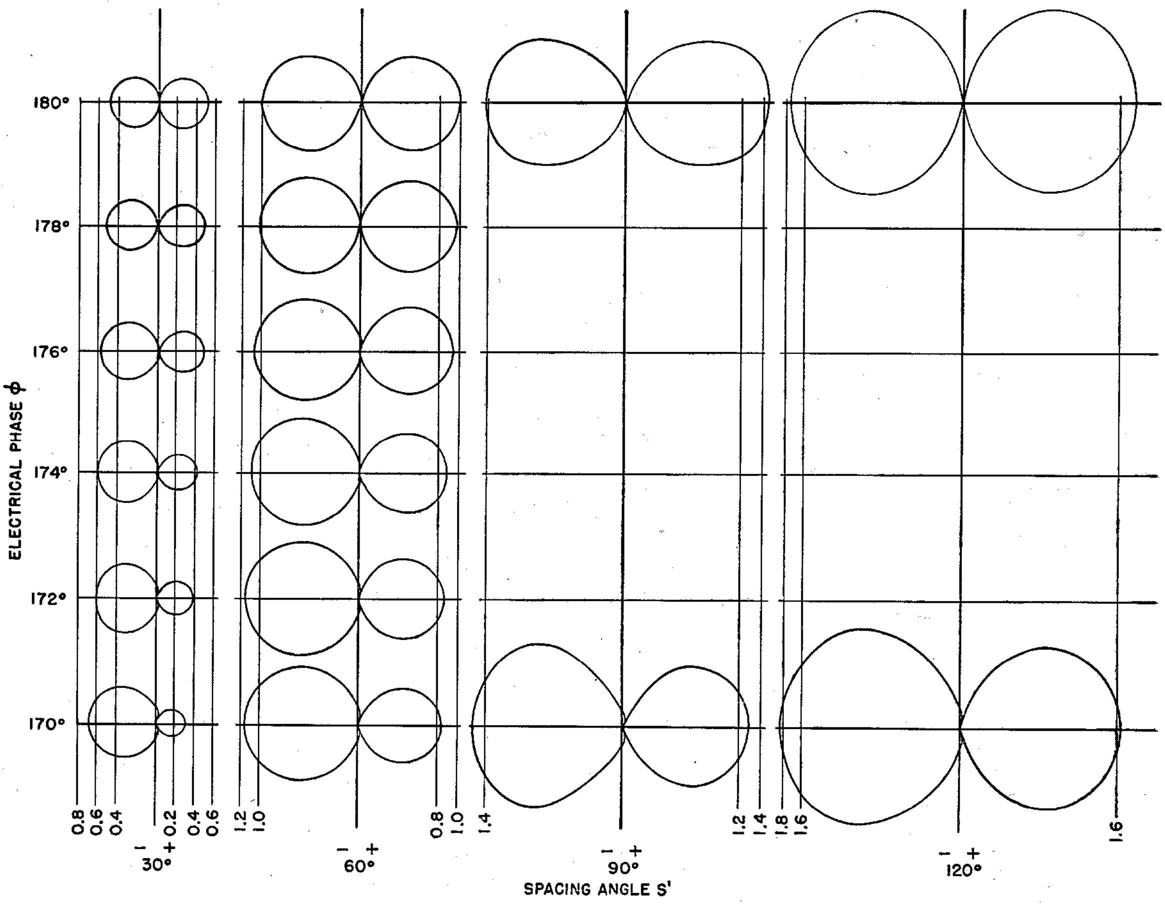

Fig. 6—Variation of field pattern with ϕ and with S'. Note the relatively large dissymmetry occurring when ϕ differs even slightly from 180 degrees.

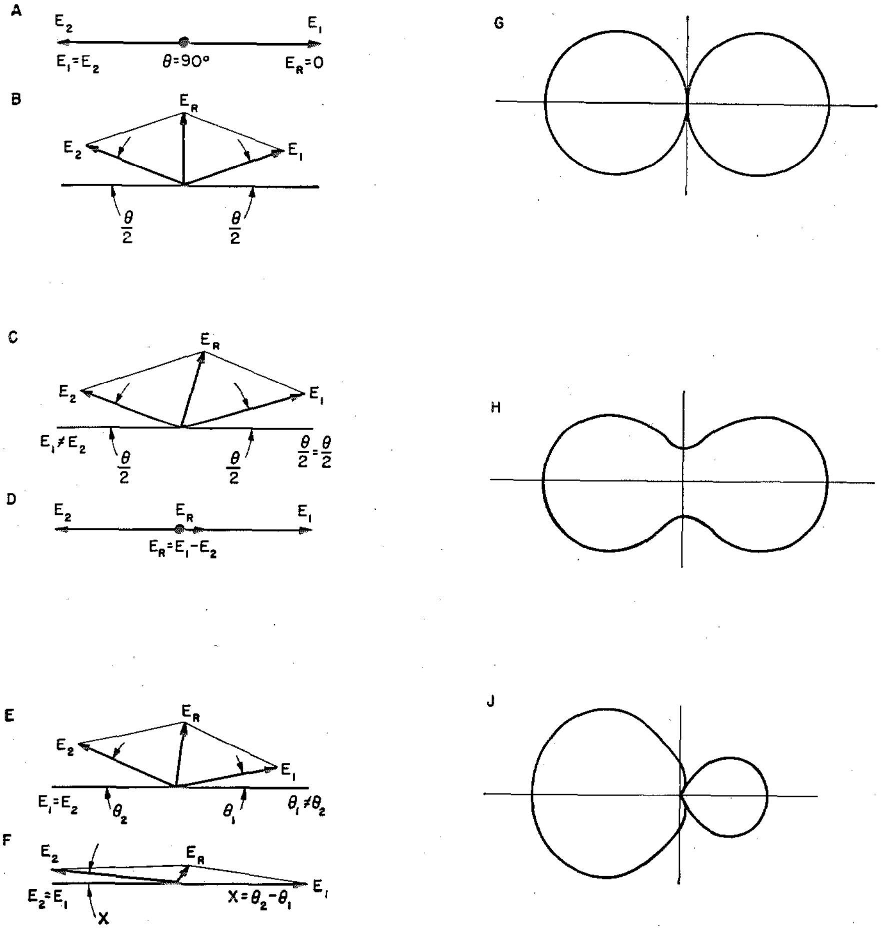

Fig. 7—Field pattern and vector diagrams illustrating various effects of the voltage and phase relationships in a directional pair array. *A, B, G*—correct figure-of-eight pattern. *C, D, H*—Pattern showing effect of amplitude inequality which results in blurred nulls. *E, F, I*—phase unbalance, causing shifted null in pattern.

not equal 180 degrees. In this case the two angles, which were shown in 7*B* and 7*C* by $\theta/2$, are not equal and have, therefore, been designated as θ_1 and θ_2. The resultant E_R is no longer at right angles to the position of E_1 and E_2 as shown in 7*A*, although its length is not greatly modified.

In Fig. 7*F*, the same condition as 7*E* has been shown except that the transmitter is at right angles to the array, i.e., at the null point. Note that E_R does not equal zero, and that it is not at right angles to the phases of E_1 and E_2.

A study of the entire Fig. 7 enables an experimental determination of whether amplitude or phase unbalance exists in a directional pair. First, if a transmitter located at right angles to the array produces a zero resultant voltage, it is certain that there is no amplitude or phase unbalance in the array. This is the condition shown at 7*A* and also at 7*B*. If there is a resultant voltage, either phase or amplitude unbalance is present.

It is quite difficult to measure the phase of the resultant voltage as shown in 7*C* and 7*D*. The equipment designer, however, can include provisions for comparing the amplitudes from the two antennas of one pair. It is also possible to move the transmitter about the null position, and if no position is found for which a good null can be secured, we may assume that there is amplitude unbalance. If it is possible to find a position where the null is good, but not at right angles to the array, it is well to investigate the null on the opposite side of the array. If these nulls are both displaced in the same direction, it is certain that there is phase unbalance in the array, i.e., that ϕ does not equal 180 degrees.

The amount of phase unbalance can be determined by reference to the development above, which shows that a shift in phase of n degrees (where n is small) results in a shift of the nulls by an equal number of degrees. If n is more than a few degrees, the amount of null shift is greatly dependent on S'. This case is hardly worth consideration because n should always be zero. If it is more than a few degrees, the resultant array will be so inaccurate as to be useless.

An example of this might be an antenna array in which poor nulls were observed at right angles to one pair. First, obtain the best amplitude balance possible. In current designs, the control unit is equipped with voltage dividers for varying the gain of the coupling units on the separate antennas; it is possible by this means to obtain amplitude balance. The target transmitter is, therefore, set up at right angles to the array and the quality of the null is measured. Then, by adjusting the voltage dividers of the two coupling units, an attempt is made to better the quality of the null. This operation balances amplitude only. After the best null has been obtained, the target transmitter is moved slightly to one side of its former position (say, toward antenna *B*), and the null is rechecked. If results are worse, a position in the opposite direction, toward antenna *A*, should be tried. The best position might, for instance, be found at 87 degrees, i.e., 3 degrees from its correct position.

The transmitter is then taken to the 270 degrees position, i.e., on the opposite side of the array. If it is found that the null, as previously, is moved toward the zero point by the same amount (3 degrees, or to 273 degrees, instead of 270 degrees), it is certain that the transmission lines connecting the pair do not cause the same phase shift. The amount of phase difference in the above example is approximately 3 degrees. In actual practice, the phase difference can be made 1 degree or less.

The polar diagrams that will be obtained for the conditions shown in Figs. 7*A* and 7*B* are shown in 7*G*. This is the perfect figure-of-eight pattern. At 7*H*, is illustrated the polar diagram which results from an inequality of amplitude. This is the condition of blurred nulls, and is illustrated vectorially at 7*C* and 7*D*. At 7*J*, is shown the polar diagram that would be obtained for the unbalance of phase illustrated vectorially at 7*E* and 7*F*. This is the shifted-null pattern.

These conditions show in a general way the results which are to be obtained from a directional pair such as is used in the spaced vertical-monopole array, as used for direction finding. (Compare with Fig. 6.)

2.3 GENERAL CONCLUSIONS

From the studies above, the following general conclusions may be drawn:

Unless k is equal to 1, E_θ can never be 0. An inequality of effective heights of antennas therefore results in blurred minima or rounded nulls.

The pattern is always symmetrical about a line formed by the intersection of the plane of the antennas with the horizontal plane.

The minima of reception occur at $\theta=\phi-90$ degrees and $\theta=90-\phi$ degrees. (A change in position of the terminal transmission line by n degrees results in a shift of the minima of n degrees, if n is small. In a current model of direction finder, the antennas of one pair are connected by an air-insulated transmission line. The characteristics of this line are such that the velocity of propagation is substantially equal to the speed of light. It may be said with considerable accuracy that the phase shift along the cable is 0.1 inch per degree per meter of wavelength. Thus, 0.1 inch is 1 degree at 1 meter, 1 inch is 1 degree at 10 meters, 10 inches are 1 degree at 100 meters, etc.)

The minimum occurs when k equals 1 and $\alpha=180$ degrees. The equation for θ under these conditions is

$$\theta=\cos^{-1}\frac{180-\phi}{S'}.$$

The two lobes of the figure-of-eight pattern are symmetrical about the minima only when ϕ is equal to 180 degrees; i.e., when the minima occur at 90 and 270 degrees.

3. *Combination of Two Directional Pairs*

Two identical directional pairs are combined at right angles to form the antenna array. The outputs from the two pairs connect separately to the orthogonal stator windings of a goniometer. It is quite evident that one stator will receive a signal proportional to the sine of θ (the angle of arrival), while the other stator receives a signal proportional to the cosine of θ. The addition of these two signal amplitudes in space quadrature creates a resultant field in the goniometer which is proportional in strength to the received signal, and which is at an angle θ. Aside from the errors of each directional pair, which were investigated in paragraph 2 above, only two sources of error resulting from distortion of the field patterns of the directional pairs are evident in the combining goniometer. The treatment following assumes that $k=1$ and $\phi=180$ degrees (paragraph 3).

3.1 QUADRANTAL ERROR

When one stator receives a stronger maximum signal from its directional pair than the other, the error is zero in the plane of either directional pair. The greatest error occurs at the 45-degree points. Therefore such errors are investigated

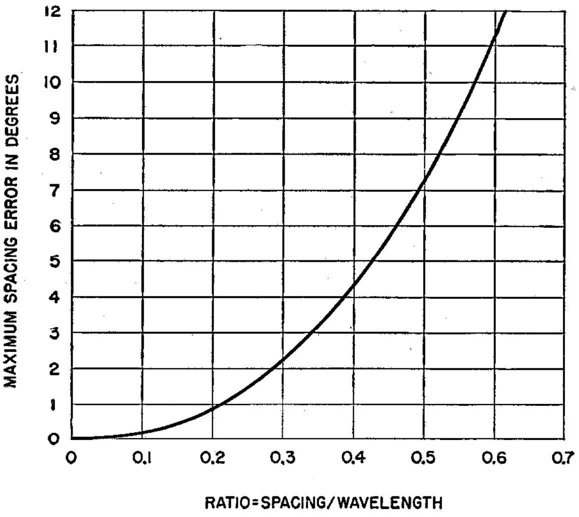

Fig. 8—Error incurred as the spacing S between elements becomes an appreciable portion of a wavelength.

experimentally by determining the error for all frequencies at the 45-degree points of the array. This is the common *quadrantal* error.

3.2 OCTANTAL ERROR

A second error occurs when the spacing S is an appreciable fraction of a wavelength. This error might have been treated under paragraph 2 as an error of the directional pair, but it is included here because its experimental determination becomes very simple when the array is studied as a whole. This error is greatest on the 22.5-degree points. It is the common spacing or *octantal* error, and has been completely and rigorously covered by various authors. Note that this error is zero at the 45-degree points, and that the value of the error is sinusoidal between its maximum (22.5-degree) and its minimum (45-degree) points.

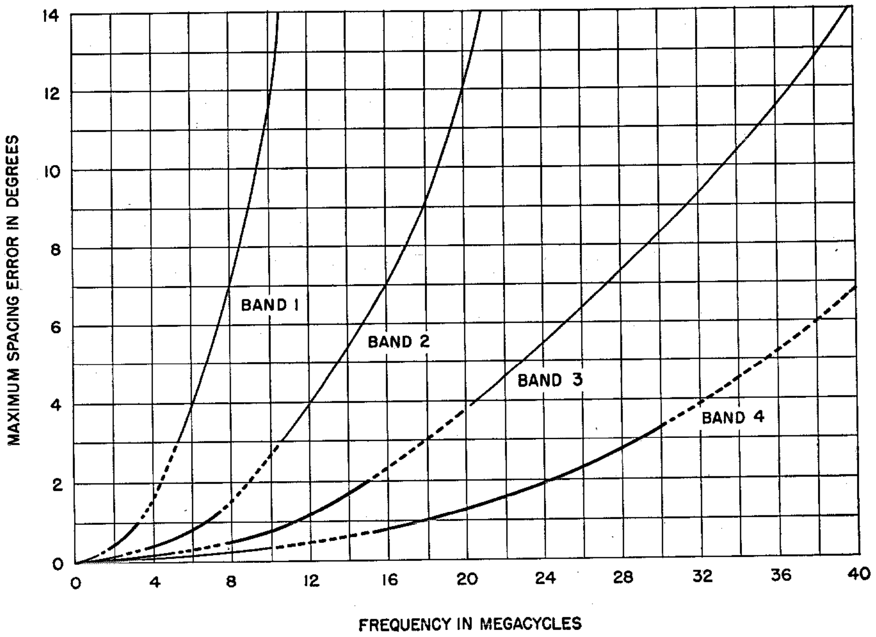

Fig. 9—Maximum spacing error in arrays covering 1.5 to 30 megacycles in four bands. The heavy solid line indicates the nominal range for each band, the dotted line indicates the extended range, and the light solid line shows errors which occur outside the range of each band.

In Fig. 8 the maximum spacing error in degrees has been plotted against the ratio of spacing to wavelength when this ratio lies between 0 and 0.6. This permits a determination of the maximum spacing error to be expected in a particular equipment at the frequencies for which the equipment was designed. These results for four arrays, covering the frequency range of 1.5 to 30 megacycles, are plotted in Fig. 9, where for each band the maximum spacing error to be expected is plotted against frequency. The maximum errors are given in Table II.

3.3 RANDOM-PATTERN DISTORTION

In Fig. 10A, are shown the resultant vectors, A and B, which result from a signal being received at an angle θ to an array of two orthogonal pairs of spaced collectors. The diagram for each of the pairs is considered to be a perfect "figure-

TABLE II

SPACING ERROR FOR FOUR ARRAYS BETWEEN 1.5 AND 30 MC.

Range	Error in Degrees			
	Band 1	Band 2	Band 3	Band 4
Nominal	1	1.3	2.1	3.5
Extended	2.8	2.8	4	7

of-eight." Therefore these two vectors, combined in the orthogonal stator windings of the goniometer, will produce a resultant at an angle θ without error.

At 10B, the same signal is shown being received at the same angle on a collector system which does not give perfect figure-of-eight reception and, therefore, the result is vectors A' and B' whose relationship to the correct vectors A and B is given by:

$$A' = XA$$
$$B' = YB.$$

In 10C is shown the resultant of the vectors A' and B' as obtained in the goniometer. This resultant is at an angle β to the correct resultant. β is therefore the correction that must be applied to the indicated bearing.

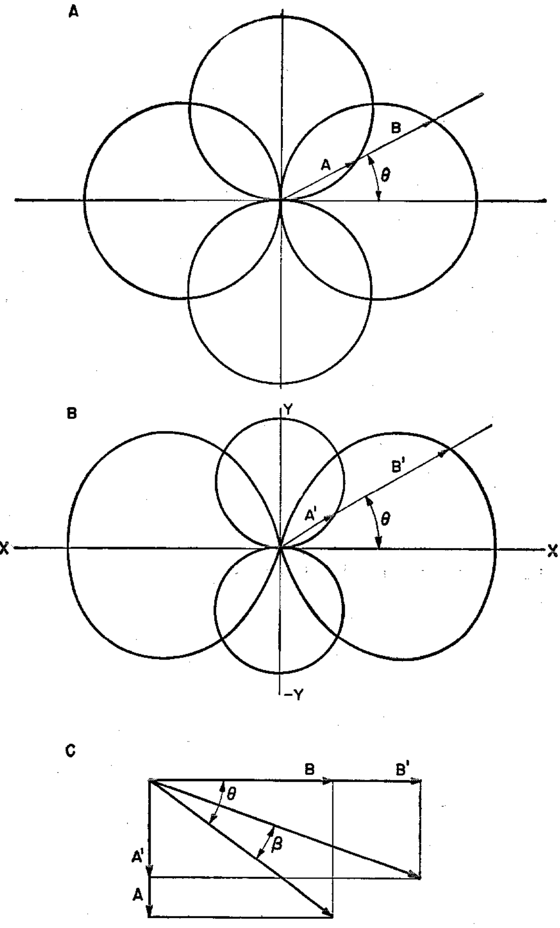

Fig. 10—A perfect set of figure-of-eight patterns are shown in *A*. In *B* the patterns are not symmetrical. The vectors in *C* illustrate the correction β which must be applied to the indicated bearing to obtain the correct bearing.

The equation for β is:

$$\beta = \tan^{-1}\left(\frac{XA}{YB}\right) - \theta. \tag{2}$$

In Figs. 11A and B are plotted values of β for various ratios of Y/X and θ in increments from 10 to 80 degrees. Note that this is a perfectly general analysis of this error. The results may be applied graphically to any system of collectors, provided their field pattern is known. By this graphic method, it is possible to postulate the error for any field configuration of either pair.

3.4 PHASE ERRORS

Phase errors occur in a system of two pairs when the signals do not arrive at the goniometer in the same phase relationship as the voltages which have been induced by the incoming wave in the corresponding array elements.

This condition will always occur when the velocities of propagation or the lengths of the cables between the two directional pairs of the array and the corresponding goniometer stators are not equal.

This results in a maximum error at the 45-degree points, and in blurred minima for signals in this direction. The error in the planes of the array pairs is zero and good minima are observed here; the maximum error is at the 45-degree points, and blurred minima or rounded nulls are observed simultaneously with the error.

In a properly designed and installed direction finder, this error is never observed, and the study will not be carried further in this paper.

4. *Goniometer*

An indicator system used in conjunction with the goniometer may consist simply of a pointer and dial arrangement attached to the shaft of the goniometer for finding an aural null or, more elaborately, of a continuously rotating goniometer together with a cathode-ray indicating device. The screen of the cathode-ray tube usually has an azimuth scale associated with it. The instantaneous position of the goniometer must then correspond exactly with the instantaneous angular deflection of the cathode-ray-tube beam. The indicator usually contains in order: the goniometer, a motor drive, and a device for controling the circular deflection of the cathode-ray-tube beam. The shafts of these three units

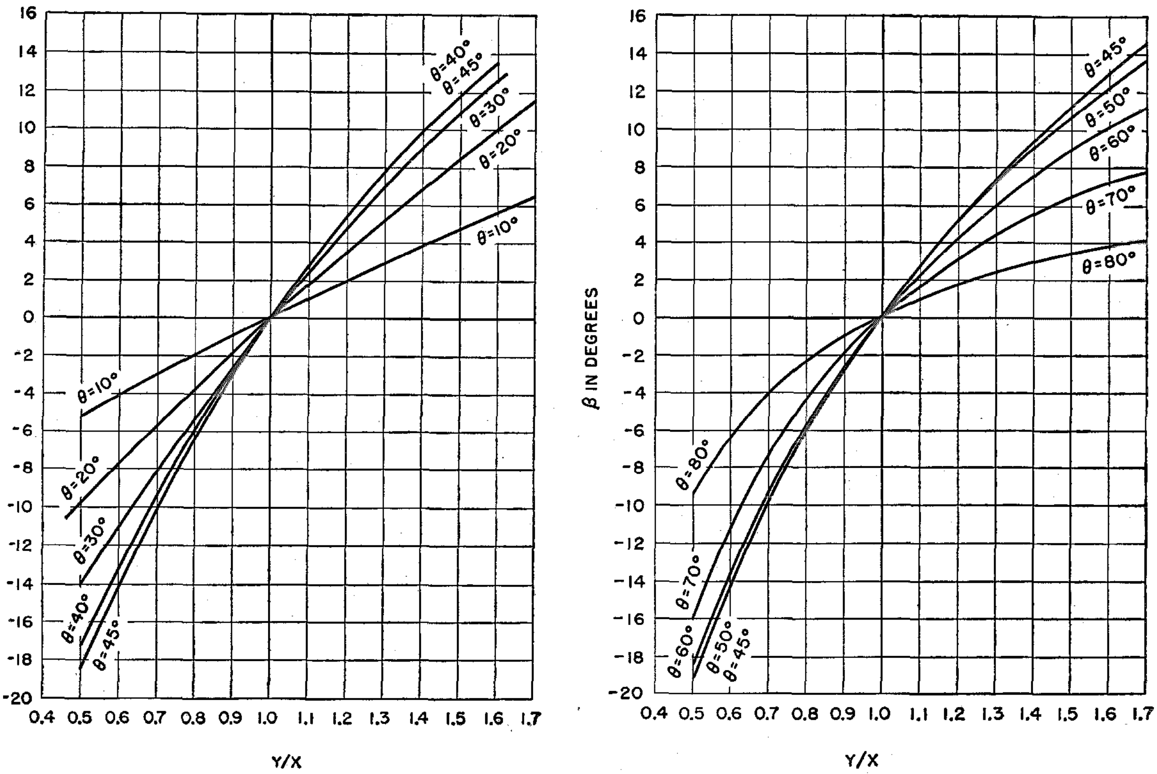

Fig. 11—Correction term β as a function of the angle of reception θ for different ratios of relative amplitude between reception from X and Y figure-of-eight patterns (Fig. 10B). At the left, for angles up to 45 degrees, the error increases, after which it again decreases as at the right.

must lie on the same straight line. Mechanical considerations and provision for maintenance usually require that the units be separable. Since it is almost impossible to keep the shafts aligned, some flexibility is provided. Initial alignment and realignment after demounting for maintenance are important considerations in maintaining the over-all accuracy of the system.

Goniometers are usually carefully designed for very small errors in the operating range. Many production goniometers have been thoroughly tested by several independent methods, and the maximum error recorded has always been less than 2 degrees.

TABLE III

ERRORS RESULTING FROM MISALIGNMENT IN GONIOMETER

Misalignment a in Inches	Pattern Bend (2α)	
	Front-Coupling Error in Minutes	Rear-Coupling Error in Minutes
0.001	4	10
0.005	21	56
0.010	42	110

The over-all equipment, however, may give errors larger than this if the goniometer is not carefully aligned both mechanically and electrically with the indicating means. The indicator used for illustration has been previously described.[3]

There are two universal joints in the indicator which might cause a bending of the pattern. The joints are between the motor and the goniometer, and between the motor and the indicator shaft. Misalignment at either of these points will cause errors. The coupling between the motor and the indicator has a fixed arm length of 1.625 inches. That between the motor and the goniometer has a fixed arm length of 0.625 inch.

If the misalignment in inches is a, and the fixed arm length is d, then the total pattern bend 2α will be given by

$$\alpha = \tan^{-1} \frac{a \sin \theta}{d - a \cos \theta}. \qquad (3)$$

The maximum value of (3) to be expected in the indicator described is shown in Table III.

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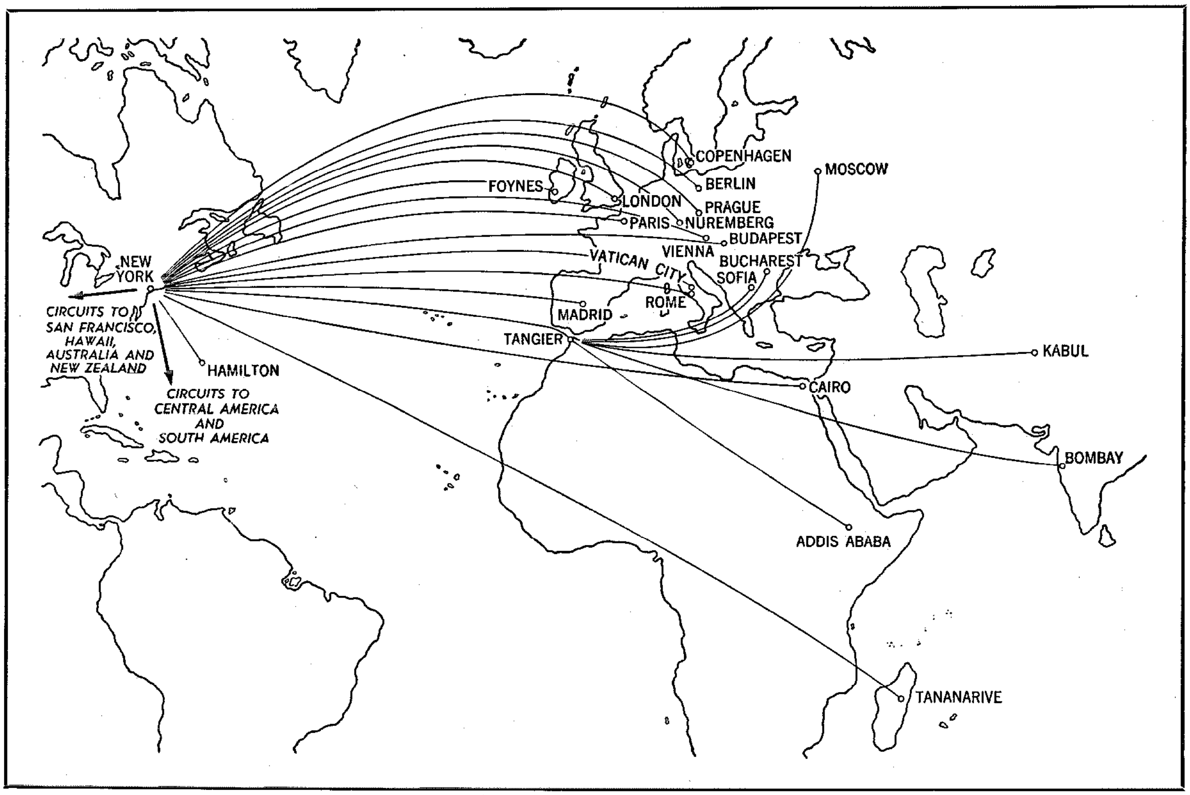
[3] Giltner Twist, "Army Radio D-F Networks," *Electronics*, v. 17, pp. 118–124; November, 1944.

Tangier Radio Relay Station

WITH THE ADVENT of technical improvements, which make practicable for radiotelegraph circuits the use of mechanical terminal equipment such as printing apparatus, automatic or semiautomatic operation of circuits in tandem becomes possible. By automatic is meant the direct relay of signals by regenerating them and removing distortions that occur during propagation. By semiautomatic is meant the reperforating of messages on paper tape and the use of this tape for retransmission.

Many radio transmission paths, such as the routes between New York and Moscow, and New York and Bombay, traverse areas near the magnetic poles of the earth. Such radio circuits either are subject to poor propagation conditions, or are so long as to offer only a limited number of hours of daily use, in addition to which they are very seriously affected by such solar disturbances as magnetic storms, aurora, and sunspots.

Mackay Radio and Telegraph Company overcame these difficulties in the Atlantic area first by establishing relay operations through its wartime station at Algiers, North Aftica, in 1944. In 1945, a permanent relay station was inaugurated in the International Zone of Tangier, North Africa. This station provides a through teletype service between New York and Moscow and furnishes twenty-four hour service from New York to other distant points such as Bombay, India; Kabul, Afghanistan; Sofia, Bulgaria; Bucharest, Rumania; and Addis Ababa, Ethiopia. The path between New York and Tangier is sufficiently distant from the north magnetic pole so that ionospheric disturbances affect it only to a minor extent. The paths between Tangier and points in Europe, Asia, and Africa are still more remote from the pole and, therefore, quite immune from such effects. The reliability of these relayed circuits is consequently of a high order.



Mackay Radio and Telegraph Company radiotelegraph circuits showing the new radio relay links operated from New York via Tangier to Moscow, Bombay, Kabul, Sofia, Bucharest, and Addis Ababa.

Carrier-Frequency Submarine Cable and Equipment for U.S.S.R.

By E. S. McLARN and G. H. GRAY

International Standard Electric Corporation, New York, New York

LESS than eight months after placement of the order, a submarine cable to connect two terminals 240 kilometers apart, terminal equipment, and one repeater were constructed, tested, packed, and readied for shipment to the U.S.S.R. Design and constructional details of the cable and terminal equipment are given.

• • •

During the last quarter of 1942, an inquiry was received from the Purchasing Commission of the U.S.S.R., through the U.S. Treasury Department, for an urgently needed submarine cable and all accessory equipment to provide four two-way telephone channels and four two-way telegraph channels between two points about 240 kilometers (150 miles) apart. An island 40 kilometers (25 miles) from one end of the cable was available for the installation of a repeater if this was considered necessary.

The proposed location of the cable was not given, but it was stated that the following conditions existed:

A. Maximum depth of water was 1300 feet (near center of longer section).

B. Static conditions very bad.

C. Salt content of the water less than that of the ocean.

D. Water at shore ends of both sections mixed with free oil, sulphur, and hydrogen.

E. Sea bottom generally smooth, but slightly stony near shore ends.

F. No currents strong enough to cause motion of cable.

G. Depth of water increases very gradually being less than 3 meters (10 feet) up to 200 meters (656 feet) from shore.

H. No power lines within 1 kilometer (5/8 mile) of proposed cable site.

Although the installation presented many new problems among which may be mentioned the great length, the number of channels required, and the fact that, because of war conditions, neither gutta percha nor paragutta were available, International Standard Electric Corporation agreed to assume the responsibility as prime contractor for the entire project up to and including the safe loading of the cable and equipment on board ship at a port in the U.S.A.

The U.S.S.R. was responsible for the safe delivery of the material to its destination as well as for the laying and subsequent testing of the cable, the installing and testing of the equipment, and putting the circuits in service.

Urgent military necessity required that the cable system be engineered and manufactured in the U.S.A., and arrangements were made through the Western Electric Company to have the Bell Telephone Laboratories undertake the over-all design of the system and the detailed design and manufacture of the terminal and repeater equipment. The Simplex Wire and Cable Company accepted the order to design and manufacture the cable under the general supervision of Bell Telephone Laboratories from the standpoints of performance and testing.

Although it was not a part of the original contract, International Standard Electric Corporation also undertook to furnish a considerable amount of information relating to cable-laying equipment and methods and to the testing of the cable after laying.

1. *Design of Cable*

A study was undertaken of possible substitutes for gutta percha and paragutta, and it was decided that deproteinized (anhydrex) rubber offered the best chance of success.

Accordingly, some 30 samples of variously constructed cables were made by Simplex Wire and Cable Company and, from tests on these samples by Bell Telephone Laboratories, the best construction was chosen. This was of the concentric type with a stranded center conductor of 10 small tinned copper wires, 0.0414 inch in diameter, laid spirally in parallel over a heavy central conductor 0.0987 inch in diameter. The total weight of the copper core was 500 pounds per nautical mile (122 kilograms per kilometer). This conductor was insulated with 1/4-inch thickness of deproteinized rubber. Over this was laid a noncorrosive rubber-filled tape which provided the base for the return conductor. This latter consisted of 48 tinned copper wires each 0.037 inch in diameter laid in parallel and applied spirally.

To protect against possible harmful effects from the crude oil in the water near the shore, about 2 kilometers of each shore end was covered by a plastic seal, 5/64 inch thick, extruded over the cable core. Two different weights of armoring were used: a light armor for the deep-sea sections and a heavier armor for the shore ends. The latter was provided both with and without the plastic seal, necessitating the manufacture of three different types of cable. The cores of all three types were identical up to the plastic seal. Layers of jute yarn and filled tape were placed throughout the cable as required. About 60 kilometers of cable, mostly of the heavily armored plastic-seal type, were provided as spare. Fig. 1 shows the layups of the three different types of cable.

The completed cable had the approximate over-all weights and outside diameters given in Table I.

Before making the approximately 500-kilometer journey from the point of manufacture to the port of loading, the cable was spliced in sections varying in length from 10 to 50 kilometers. Lengths too great for one flat-car were coiled on several cars, with a loop of cable between cars to prevent cable injury. For the 50-kilometer lengths, 5 cars were required.

By June, 1943, less than 6 months from the date the order was placed, the entire cable had been manufactured, tested, and shipped to the steamer.

At a temperature of 24 degrees centigrade (75 degrees Fahrenheit), the direct-current resistance of the central conductor averaged 1.214 ohms and that of the return conductor 0.532 ohm per kilometer.

The characteristic impedance over the range used for the telephone channels, about 3.2 kilocycles to slightly under 30 kilocycles, was fairly constant at about 54 ohms. Over the telegraph range from 300 to 3000 cycles the impedance was somewhat higher.

To determine the attenuation of a section of the cable, one end was terminated in its image impedance and a frequency of 40 kilocycles applied at the other end. The currents at the two ends were determined by thermocouples. With the 40-kilocycle frequency replaced by direct current and the cable by a calibrated resistance box, the input voltage and the resistance box were adjusted until the two thermocouples gave the same indications as before. The attenuation was then computed from the value of the box resistance and the image resistance of the cable.

The attenuation at 40 kilocycles and 52 degrees centigrade (126 degrees Fahrenheit) as measured under water in tanks at the factory varied from 1.00 to 1.04 decibels per nautical mile and averaged about 1.02 decibels per nautical mile or 0.55 decibel per kilometer. The guaranteed maximum value of attenuation was 1.13 decibels per nautical mile.

TABLE I

Physcial Characteristics of Cable

Type of Cable	Weight		Outside Diameter	
	Pounds per Nautical Mile	Kilograms per Kilometer	Inches	Millimeters
Light Armor, No Oil Seal	15,200	3750	1.55	40.25
Heavy Armor, No Oil Seal	22,000	5400	1.84	46.75
Heavy Armor, Oil Seal	25,750	6310	2.01	51.05

It is interesting to note that the cable not only met the specification requirements in every respect but its over-all transmission capabilities proved to be approximately 10 percent better than were specified.

2. *Carrier Equipment—General*

As four two-way telephone and four two-way telegraph channels were required, carrier offered the only means of furnishing these circuits over one cable. It was decided that an intermediate repeater would be necessary on the island previously mentioned. In addition to the commercial channels, a telegraph order wire was provided from each terminal to the repeater station. This could be used for through telegraph communication (terminal to terminal) by connecting the direct-current extension circuits of the order wires at the repeater station. Teletypewriter equipment was installed for use on the order wire.

For convenience, the terminal at the end of the 200-kilometer section of cable is hereafter referred to as *A*, that at the end of the 40-kilometer section as *C*, and the repeater at the junction of the two as *B*.

To meet the quick delivery required, it was decided to use standard panels, or parts of panels, to as great an extent as possible, even though this did not result in quite as satisfactory an arrangement as could have been obtained from an entirely new design. One result of this decision was that the equipment made use of a larger number of different types of vacuum tubes than would ordinarily have been necessary.

The frequency arrangement selected is shown in Table II.

TABLE II

Direction of Transmission	Circuit Number	Frequency in Cycles per Second			
		Order Wire	Telegraph Channels	Telephone Channels	
				Carrier	Pass Band
B-A	—	255	—	—	—
C-A	1	—	425	5,900	3,200– 5,600
C-A	2	—	595	8,850	6,150– 8,550
C-A	3	—	765	11,800	9,100–11,500
C-A	4	—	935	14,750	12,050–14,450
C-B	—	1105	—	—	—
A-C	4	—	1615	17,700	18,000–20,400
A-C	3	—	1785	20,650	20,950–23,350
A-C	2	—	1955	23,600	23,900–26,300
A-C	1	—	2125	26,550	26,850–29,250
A-B	—	2295	—	—	—
B-C	—	2465	—	—	—
				32,450 (Group Modulator and Demodulator)	

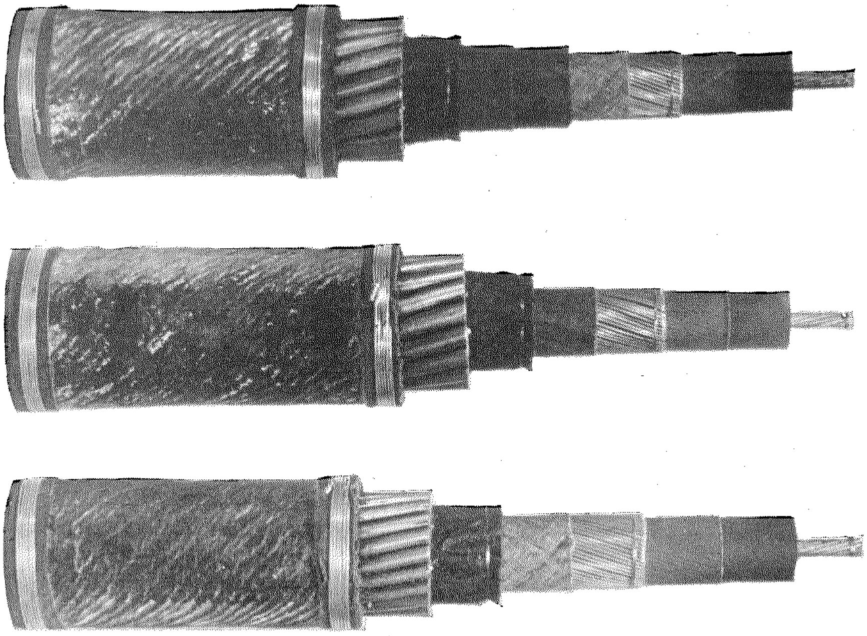


Fig. 1—Three types of cable were used and differed in the weight of armoring and in the application of an oil seal for the shallow-water sections. All have identical center conductors, 1/4-inch wall of anhydrex insulation, 0.012-inch layer of noncorrosive rubber-filled tape, coaxial conductor of 48 strands of round wire, and a 0.012-inch noncorrosive rubber-filled tape. The top cable has a 5/64-inch Plastex oil seal and another 0.012-inch rubber-filled tape. The two lower cables do not have an oil seal. All three then have a serving of 200-pound cutched jute yarn, galvanized steel armor wires, and two serves of impregnated jute yarn. The outside diameters in inches are 2.01, 1.84, and 1.545, respectively.

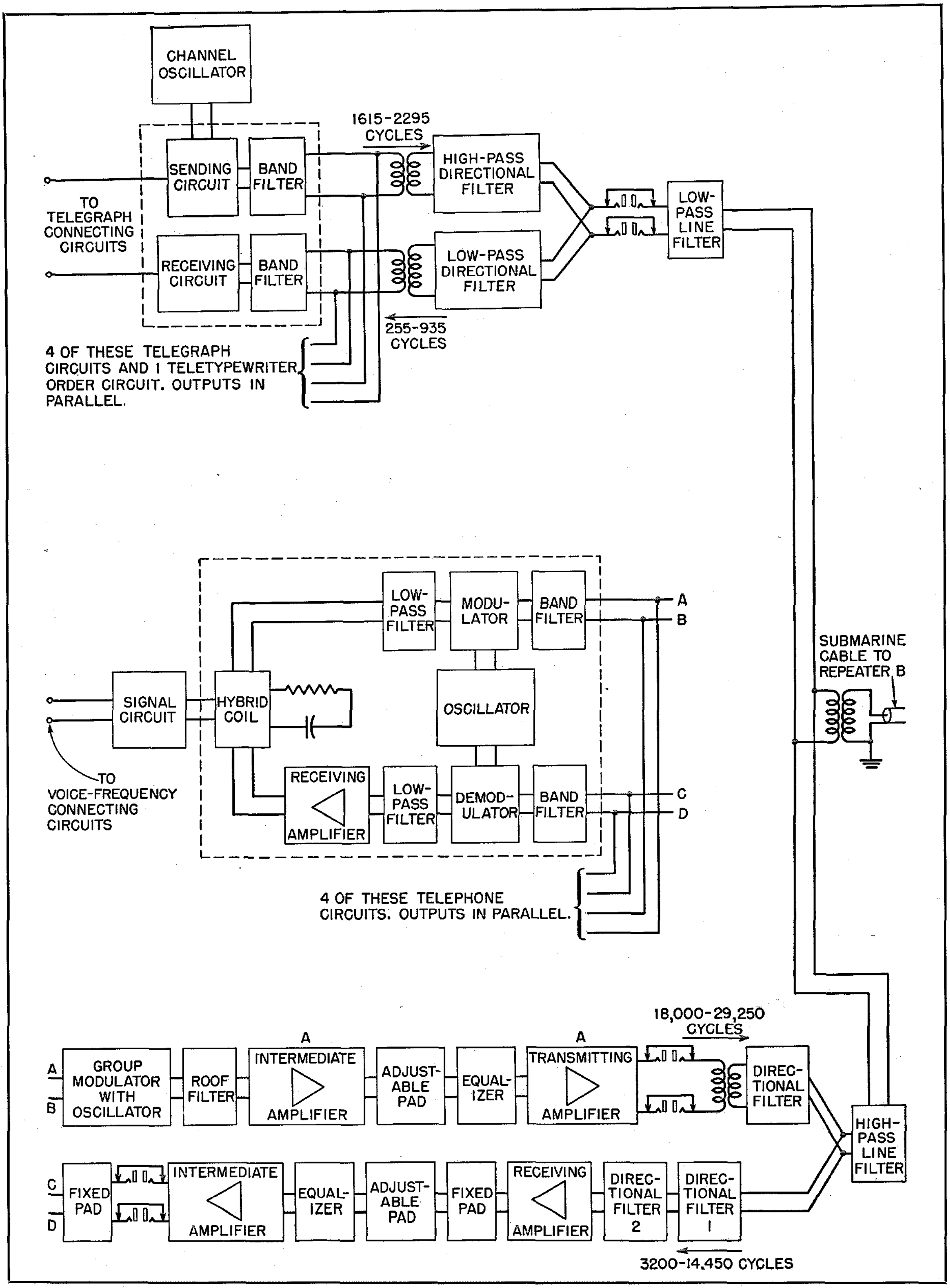


Fig. 2—Schematic arrangement of the carrier telephone and telegraph equipment; Terminal *A* equipment.

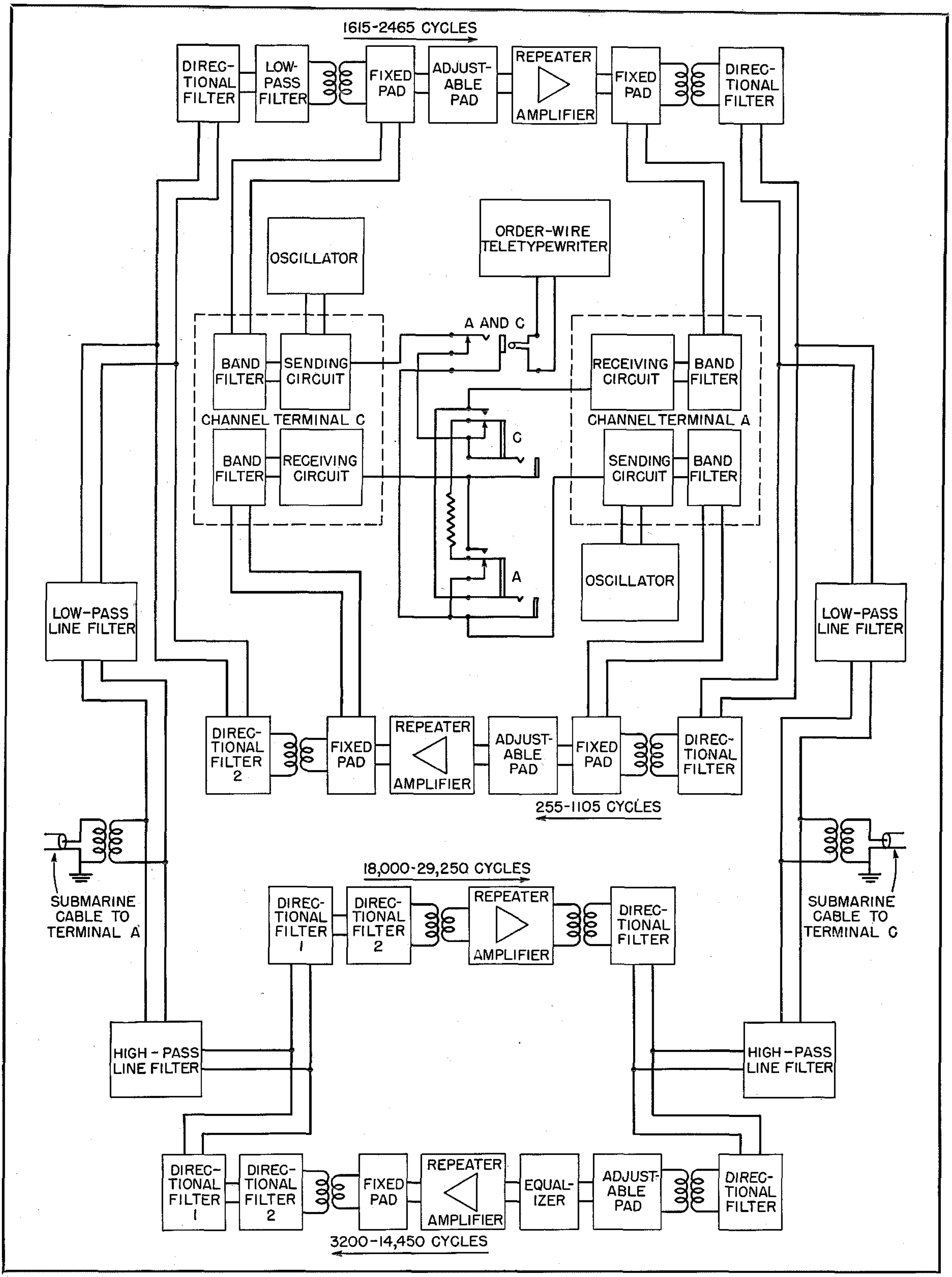


Fig. 3—Schematic arrangement of Repeater *B* equipment.

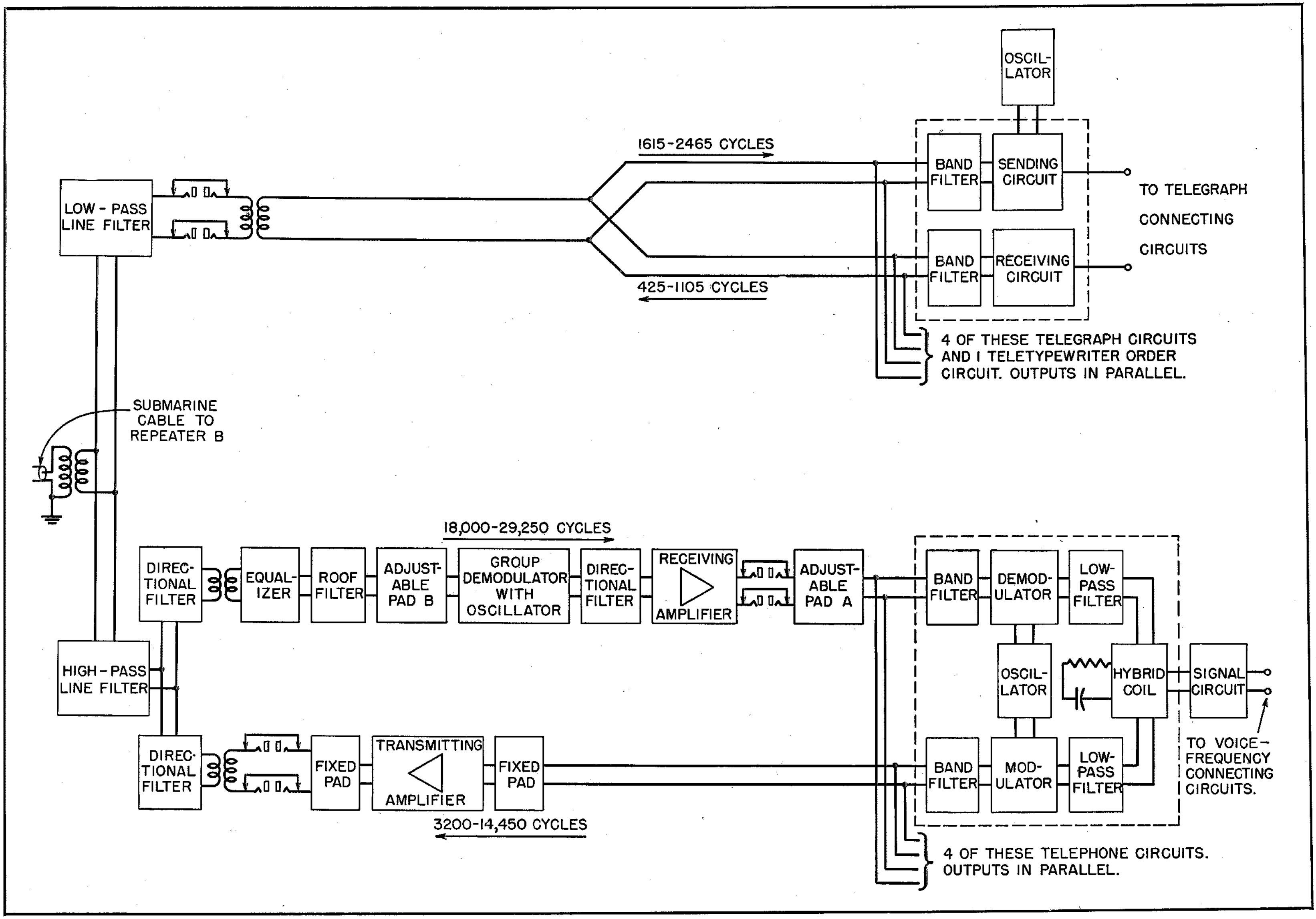


Fig. 4—Schematic arrangement of Terminal *C* equipment.

It will be noted that each telephone channel transmitted a band extending from about 300 cycles to 2700 cycles. Single-sideband transmission was used with suppressed carrier. The telegraph system transmitted the carrier and both sidebands.

Separate terminal and repeater equipment was used for the telephone and telegraph circuits, the separation being made by low-pass and high-pass line filters. A common power supply was, however, used for all equipment at each location. To obtain the proper impedance match, a 600- to 54-ohm repeating coil was installed at each end of each cable section.

Figs. 2, 3, and 4 show over-all application schematics of both terminals and the repeater station. The general arrangement is clear from these figures but the following comments are made as to details.

A. Copper-oxide modulators and demodulators were used.

B. The intermediate and transmitting amplifiers shown in the "high group" transmitting path of terminal *A* are capable of furnishing a gain of 53 decibels each.

C. The amplifier in the "high group" transmitting path at the repeater *B* is capable of a gain of 66 decibels.

D. The receiving amplifier provides a fixed gain of 50 decibels.

E. Lower gains are, of course, satisfactory for transmitting in the reverse, *C-A* direction, as the line losses are smaller.

F. 1000 cycles, interrupted at a 20-cycle rate, is used for ringing.

G. It is expected that each channel will operate at an over-all loss of about 9 decibels.

H. The sending circuit of each telegraph channel includes means for interrupting the carrier in accordance with the signals to be sent.

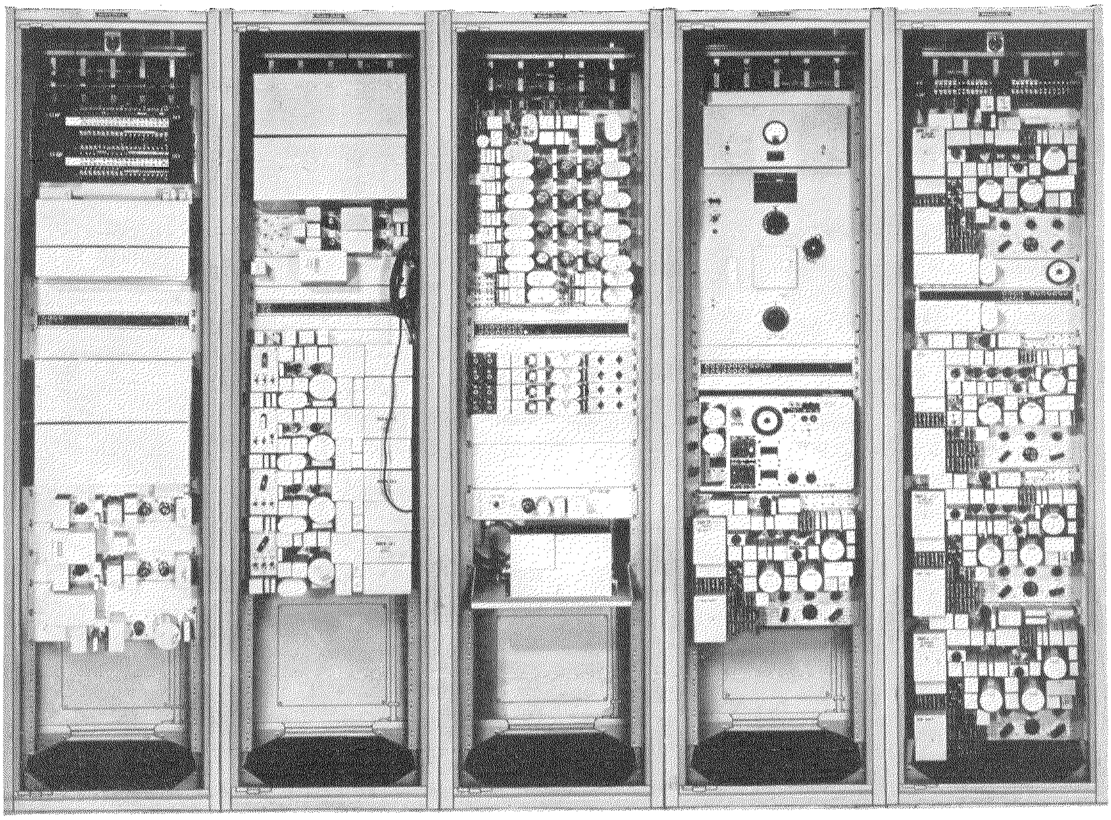


Fig. 5—Telephone and telegraph bays for terminal *A* with doors removed. From left to right are the following bays: telephone high group, telephone low group and channel, signaling, test and order wire, and telegraph.

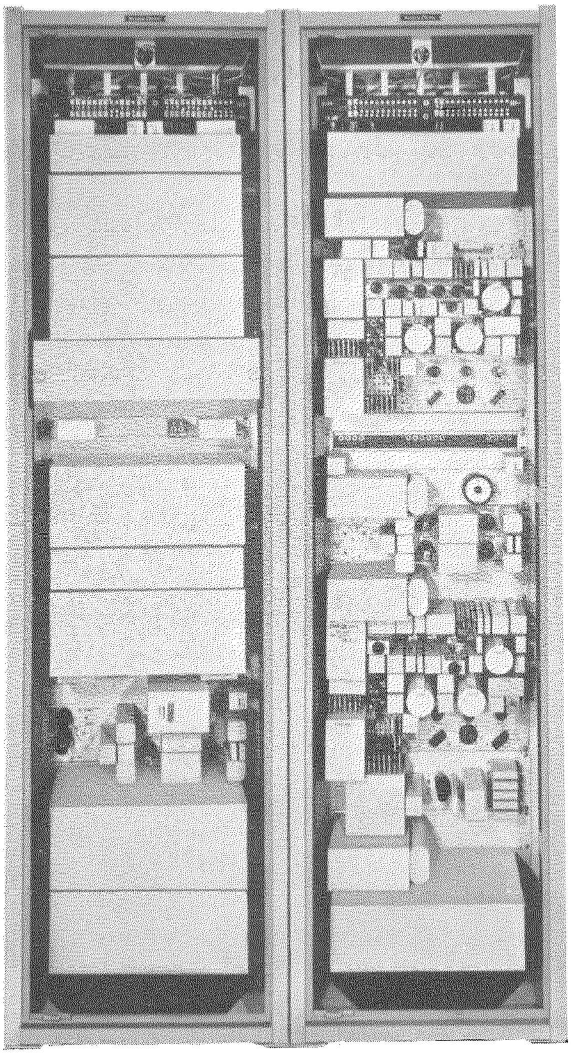


Fig. 6—Telephone (left) and telegraph bays of the repeater equipment with doors removed.

I. The telegraph terminals were designed for use with different types of direct-current extension circuits or "loops," such as:

Neutral to Positive Battery—Half and Full Duplex
Neutral to Negative Battery—Half and Full Duplex
Polar Full Duplex.

J. Teletypewriters may be connected directly to the telegraph terminal or through telegraph switchboards or repeaters.

K. The jacks at *B* were so arranged that teletypewriter communication could be carried on with both terminals simultaneously or with either terminal separately.

Fig. 5 gives a general idea of the arrangement of the equipment in the cabinets of terminal *A*, and Fig. 6 is a similar view of the repeater equipment. Terminal *C* is very much like terminal *A* but has slightly less equipment.

3. Power Supply

Both terminals were designed to operate normally from 110–125-volt, 50/60-cycle mains. For emergency use, a gas-engine-driven alternator was provided. At the repeater station, two gas-engine sets were furnished as there was no commercial supply. Regulated tube rectifiers, operating from the alternating-current mains, or from the alternators, were used to provide the direct voltages, the following being required at each terminal: $+150$, $+130$, -130 and -24. At the repeater point, direct voltages of $+150$, $+65$, -65 and -24 were provided.

Each regular direct-voltage supply consisted of a full-wave rectifier with an associated battery to act as a filter, take care of peak loads, and provide a short reserve in case of a power failure. Suitable alarms provided indications of high- and low-voltage conditions or of failure of any source of power.

Static ringing generators converted the input of 105–125 volts at 50 or 60 cycles to 75–100 volts at $16\frac{2}{3}$ or 20 cycles which were used for the normal 20-cycle signaling supply. On failure of the alternating-current source, transfer was automatically provided to a 24-volt battery-driven generator.

The general arrangement of the power equipment at the *A* or *C* terminal is shown by Fig. 7. The power equipment for the repeater is shown in Fig. 8.

4. Testing Facilities

For testing purposes, oscillators, transmission-measuring sets, current-flow test sets, voltohmmeters, relay test sets, and vacuum-tube test sets were provided.

5. Mounting Arrangements

The equipment was mounted in steel cabinets approximately 7 feet by $22\frac{1}{4}$ inches by 17 inches, with the exception of those for the 24-volt power equipment which were 4 inches wider.

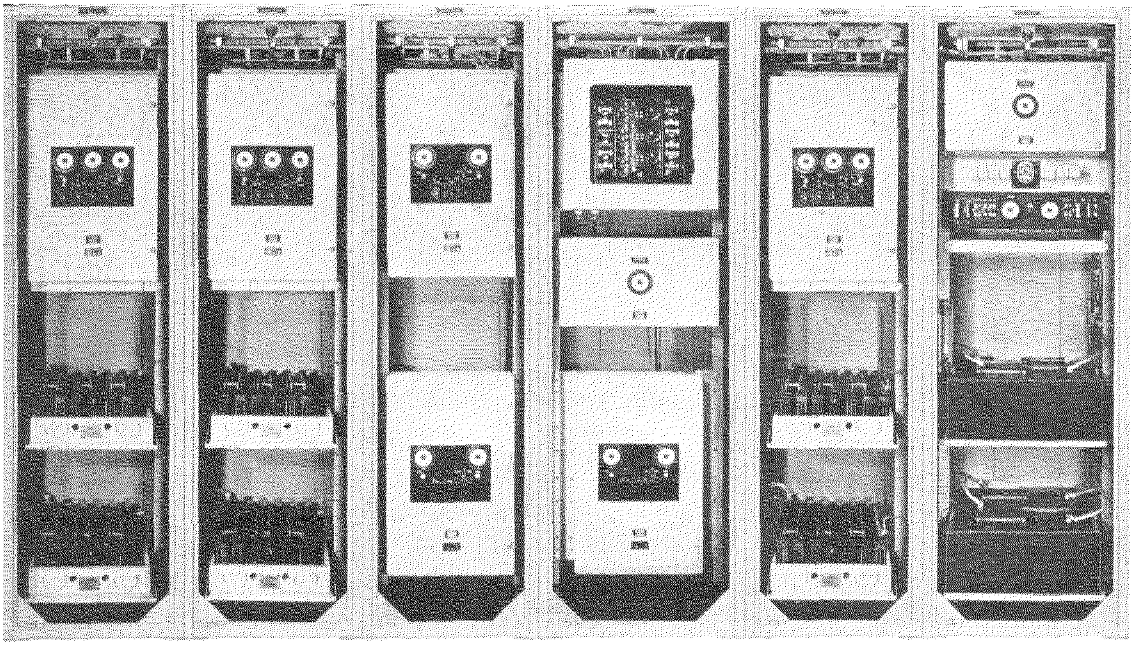


Fig. 7—Power supply equipments for the *A* or *C* terminals. From left to right are bays for +130 volts, −130 volts, +130-volt and −130-volt spare rectifiers, power service panel with cover removed and −24- and +150-volt spare rectifiers, +150 volts, and −24 volts. This is a front view with doors removed.

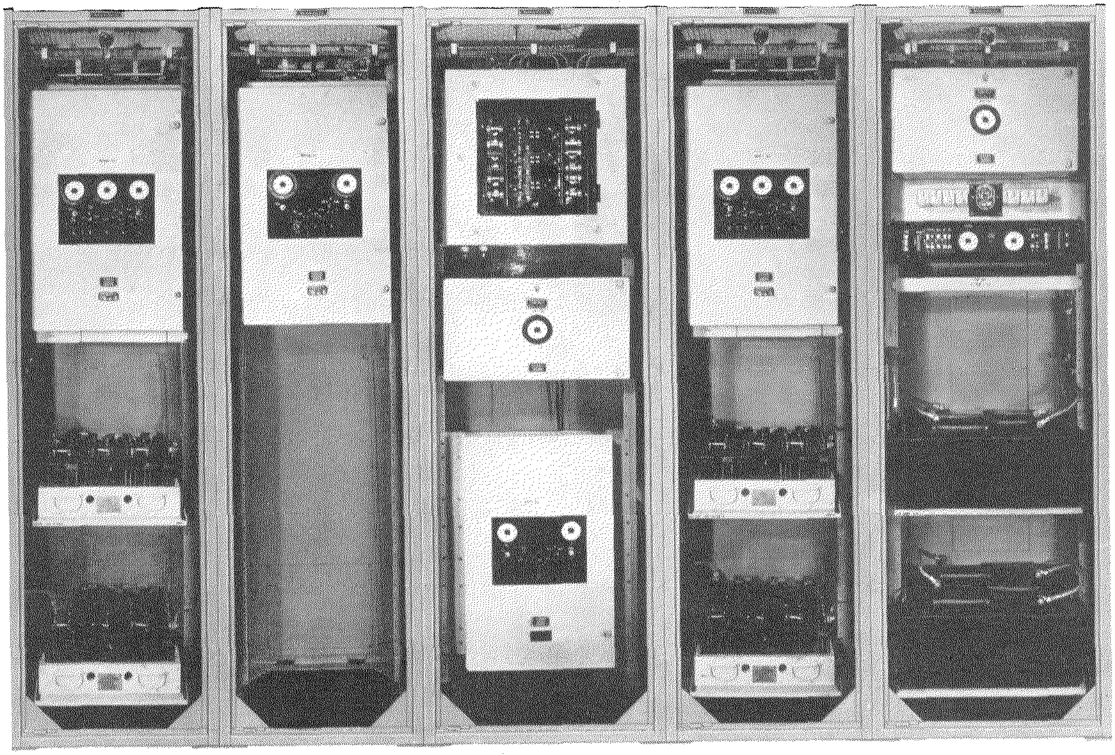


Fig. 8—Front view with doors removed of the repeater power-supply equipment. From the left are bays for ±65 volts, ±65-volt spare rectifier, power service panel with cover removed and −24-volt and +150-volt spare rectifier, +150 volts, and −24 volts.

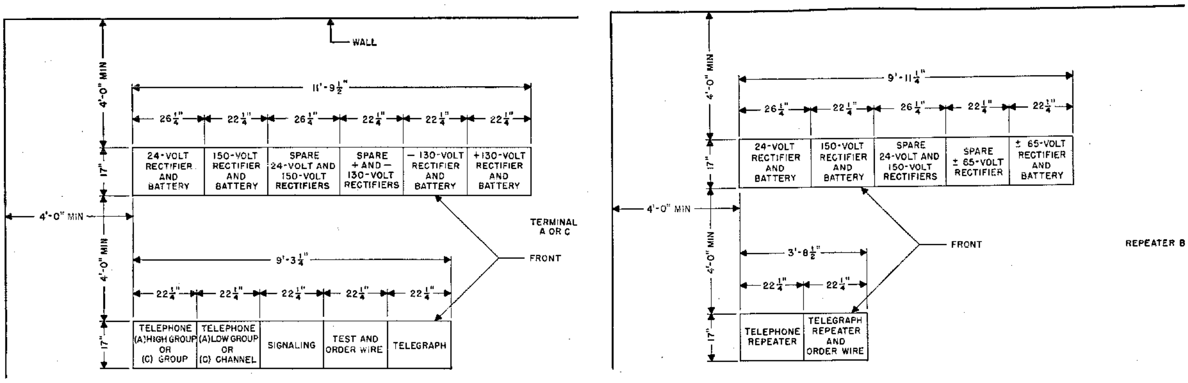

WALL
11'-9½"
26¼"
22¼"
26¼"
22¼"
22¼"
22¼"
24-VOLT RECTIFIER AND BATTERY
150-VOLT RECTIFIER AND BATTERY
SPARE 24-VOLT AND 150-VOLT RECTIFIERS
SPARE + AND − 130-VOLT RECTIFIERS
− 130-VOLT RECTIFIER AND BATTERY
+130-VOLT RECTIFIER AND BATTERY
4'-0" MIN
17"
4'-0" MIN
TERMINAL A OR C
FRONT
9'-3¼"
22¼"
22¼"
22¼"
22¼"
22¼"
17"
TELEPHONE (A) HIGH GROUP OR (C) GROUP
TELEPHONE (A) LOW GROUP OR (C) CHANNEL
SIGNALING
TEST AND ORDER WIRE
TELEGRAPH
4'-0" MIN
9'-11¼"
26¼"
22¼"
26¼"
22¼"
22¼"
17"
24-VOLT RECTIFIER AND BATTERY
150-VOLT RECTIFIER AND BATTERY
SPARE 24-VOLT AND 150-VOLT RECTIFIERS
SPARE ± 65-VOLT RECTIFIER
± 65-VOLT RECTIFIER AND BATTERY
4'-0" MIN
FRONT
REPEATER B
4'-0" MIN
3'-8½"
22¼"
22¼"
17"
TELEPHONE REPEATER
TELEGRAPH REPEATER AND ORDER WIRE

Fig. 9—Possible floor plan of the terminals and of the repeater.

All cabinets had full-length front and rear doors, each containing ventilating louvres.

A telephone handset was mounted in the telephone channel cabinet at each terminal.

Fig. 9 shows a suggested floor plan for the terminal and repeater stations and gives an idea of the amount of space required.

6. *Over-All Test*

On completion of the equipment, it was set up in the Bell Telephone Laboratories and artificial cables of the proper constants were connected between each terminal and the repeater equipment. A complete series of talking, telegraphing, and signaling tests was run. Careful measurements indicated that the desired transmission levels could very easily be obtained, that the necessary cross-talk and telegraph-interference limits had been met, and that all the signaling equipment functioned properly. It was even found possible to operate a 0-decibel circuit over the four telephone channels connected in tandem although under this condition the margin was very small. With a 9-decibel circuit loss and all four channels in tandem, a very satisfactory circuit was obtained.

The equipment was packed and ready for shipment about August 1, 1943, less than eight months from the date the order was placed.

Survey of Radio Navigational Aids

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Editor's Note: The field of radio aids to air navigation has become increasingly complex, particularly as a result of war activities. Those actively engaged in specialized aspects within the field, as well as newcomers, find it difficult to obtain a broad and up-to-date view of even the major developments and proposals. A number of co-workers suggested to the author that he prepare a compact survey of the field; the reception which this timely paper met prompted its publication in Electrical Communication.

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1. Introduction

AIRPLANE NAVIGATORS dream of a day of no sextant, no compass, no Greenwich time, no lines of position, no pressure-pattern or assorted other charts and computers, no mathematics, and no time out at the plotting board. Commerical airline operators dream of a day of no navigators.

They await the perfection of some sort of completely automatic direct-reading indicator of position, such as the latitude-longitude meter sketched in Fig. 1; or a mechanism that moves a stylus or spot of light to pinpoint the observer's position on a map; or a combination radio beam and automatic pilot that guides the airplane as unerringly along a prescribed path as the flanged wheels of a railway car. Such devices would be a great convenience, and with the rapidly increasing density, range, and speed of traffic, they become urgently desirable for aircraft use.

Many promising and glamorous radio aids for navigation have been developed or proposed for short- or medium-distance applications; radar techniques are often employed. The very long distance or transoceanic air navigational problem appears to be a knottier one. Appreciating the technical difficulties, if not sheer impossibility, such an instrument as pictured in Fig. 1 can now be called only an "artist's conception," a fantasy. But men have dreamed before, as of talking over wires, flying through the air, and using atomic energy.

With the relaxation of wartime restrictions, a number of new types of navigational aids have been disclosed, their principles no longer dissimulated by the military code names under which they masqueraded. The time is, therefore, opportune for making a general survey and classification of radio aids to navigation to see what they might contribute toward the "impossible," yet eagerly awaited, ideal system.

1.1 Dead Reckoning Versus Radio Methods

Devices like the so-called air-position indicators (API) or ground-position indicators (GPI) should be mentioned at the outset. They are in various stages of development; their indicators in fact resemble that of Fig. 1 quite closely. These are essentially dead-reckoning devices which attempt to indicate present position by computing the net direction and distance made good from some known past position of the craft. They are analogues of the "dead-reckoning tracers" sometimes used on surface ships, in which the track of the ship is mechanically plotted on a map by taking continuous account of the direction of the ship's heading as indicated by a compass, and the distance or speed of travel as indicated by the revolutions of the propeller.

The basic difficulty with such methods is that a ship or airplane is moving through a medium—the sea or the atmosphere—which is itself moving with respect to the earth. That movement of ocean current or wind drift is erratic in speed and direction, and its measurement by an observer immersed in the very medium is quite a feat. Inherently difficult problems arise in the accurate and continuous determination of true direction and distance from compass, time, acceleration, and air-speed readings, since compass variations, wind speed, wind direction, and other troublesome factors must be both determined and taken into account.

Besides, for reliability, it is fundamentally desirable that a navigational system should place no dependence on unbroken continuity of all past readings, even if all the factors were capable of being measured accurately. One should rather be able to make an independent or fresh position determination at any time instantaneously, or nearly so, "starting from scratch" so to speak. Indeed, reliability, independence, and capability of covering all traveled areas at all times are basic requirements for an ideal navigational system, for which mere convenience should be sacrificed, if necessary.

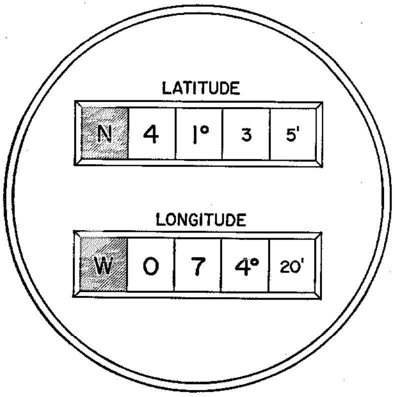

Fig. 1—An ideal indicator for navigation.

Radio waves travel great distances almost instantly; the idea of using radio for giving navigational guidance quickly and freshly occurred to people only shortly after radio proved useful for communication purposes. Many and various, even confusing, are the radio navigational schemes and their names. To evaluate them intelligently, one should have in mind a clear classification of their principles and their inherent capabilities and drawbacks; very often the differences between systems are merely in details that may be changed. Fundamental principles and characteristics are emphasized in the following description of the four basic types into which all radio navigational aids may be classified, even those whose fanciful *noms de guerre* may still only be whispered.

The classification used in this survey and the order of presentation are to a certain extent arbitrary for there are various ways of classifying radio navigational aids, such as by navigational techniques, electronic techniques, radio-frequency bands, modulation, etc. No compartmentation can be absolutely clean-cut. The simple one used, however, is quite logical and illuminating, and is a convenient framework for the survey.

I. Basic Types of Radio Navigational Systems

2. *Type A: Directional Receiving Systems*

2.1 General Principle

These are the oldest radio navigational aids and probably will always be useful to some extent. Their principle is based on the fact that certain types of antennas, loops in the simplest case, receive radio energy strongly or weakly depending on how they are oriented with respect to the incoming direction of the radio waves. Thus, by turning the antenna and observing how the received signal strength varies, the direction from which a radio ray from a distant transmitter strikes the antenna may be determined.

Many aircraft and most surface craft carry direction finders (DF) working on this principle, such as simple radio compasses, aural-null direction finders, or automatic visual direction finders. Indications may be of the aural, meter, or cathode-ray-tube type. Some of the companies that have contributed to the development of direction finders for air and marine use include Sperry, Bendix, Radio Corporation of America, Federal, British Marconi, Société Française Radio, and a number of others. The antennas are commonly loops; but may be other special arrays such as Adcocks, horns, or curved surfaces as in radar applications.

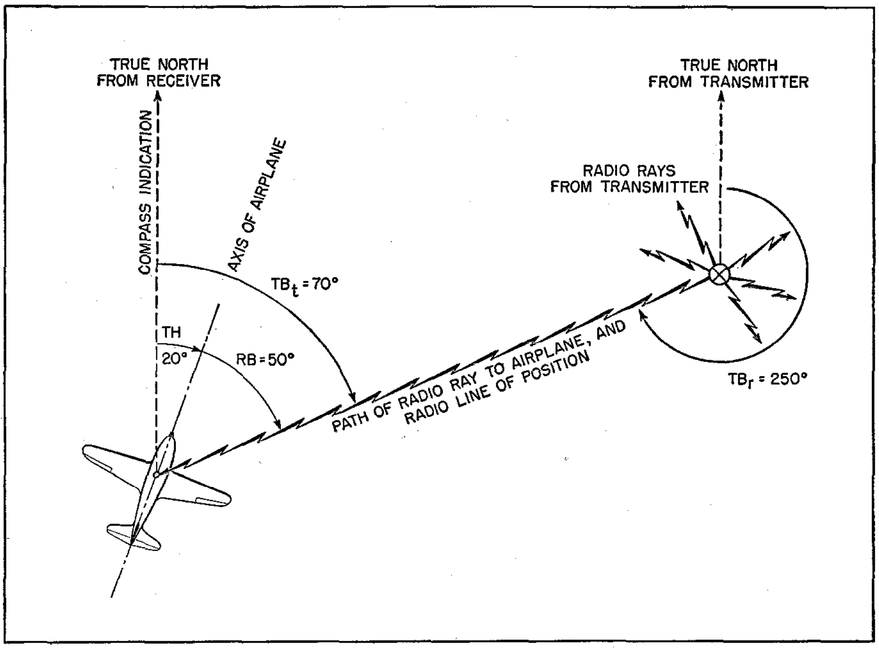
TRUE NORTH FROM RECEIVER
COMPASS INDICATION
AXIS OF AIRPLANE
TB$_t$ = 70°
TH
20°
RB = 50°
PATH OF RADIO RAY TO AIRPLANE, AND RADIO LINE OF POSITION
TRUE NORTH FROM TRANSMITTER
RADIO RAYS FROM TRANSMITTER
TB$_r$ = 250°

Fig. 2—Principle of airborne direction finding.

TH = true heading, determined by compass indication with corrections.
RB = relative bearing of the transmitter, obtained with the direction finder.
TB_t = true bearing of the transmitter = $TH + RB$.
TB_r = true bearing of the receiver = $TB_t + 180$ degrees = reciprocal of TB_t.

A great advantage is that the system always (in principle at least, neglecting errors) indicates the direction of the radio waves regardless of the type of transmitter. Its frequency, modulation, directivity, etc., is practically of no consequence; no special transmitting stations need be erected. A craft equipped with a simple direction finder can take bearings of any radiotelegraph, radiotelephone, or radio-range station within its distance and tuning range. Unfortunately, no direction finder is a completely independent or self-sufficient navigational aid, as next explained.

2.2 Airborne Direction Finding

Fig. 2 illustrates how a direction-finder reading observed on an airplane reveals the direction of the incoming waves only with respect to the airplane's fore-and-aft axis, a rather vague reference except for homing purposes.

Homing makes use of the fact that the radio ray follows the most direct or shortest path between the transmitter and receiver. This path is actually the arc of a great circle on the earth; for short distances it is practically a straight line. Thus the direction finder may be said to "point toward the transmitting station." A homing flight, however, making use of this property alone for guidance, will in general not result in the airplane actually following such a direct path. Just as in the case of courses steered by reference to magnetic compass directions or bearings, cross winds alter the actual track of the airplane into something other than a direct path.[1, 2]

To determine the *true bearing* of the radio ray, that is, its direction with respect to the local meridian or true north line, the navigator would have to read, in addition, a magnetic compass (and correct it for variation and deviation). This additional step is necessary to establish the observer's latitude-longitude position by direction-finding methods. Homing ability is only one item in navigation; ability to make a complete "fix" at any time is the basic navigational requirement.

Whether observations are aural, visual, celestial, or radio type, a fix is always established by measurements of distances, directions, or both, of the observer from other objects whose geographical positions are known. As a surface has length and breadth, any single observation locates one only partially on it. Knowledge of the distance from a known landmark means only that one is somewhere on a circle of that radius centered about the landmark. Knowledge of the direction from a known landmark means only that one is somewhere on a straight line, or great circle, drawn at that bearing from the landmark. A line of these or other shapes, along which an observer must be definitely located as a result of a single navigational observation, is called a line of position (LOP). For a complete location or fix, one must know the position of a landmark or two, make two observations, and draw *two* lines of position. The observer's position, or fix, is then definitely at the intersection (or at one of the intersections) of the two lines of position.

For short distances, say up to 200 miles or so where the earth's surface may be considered plane, this process is fairly straightforward using an airborne direction finder and a compass. Fig. 2 illustrates how one line of position may be drawn. Note that the direction finder and compass combination gives the direction of the transmitter as "seen" from the airplane, position unknown. A direction from an unknown point is rather vague information. To draw a line of position, the direction of the airplane from the position of the transmitter must be known. It is obvious that if a person on an airplane were to sight (by eye or by radio) along a straight line to the transmitting station, then a person at the transmitter would have to sight exactly backwards along the same straight line to see the airplane. And, as assumed in Fig. 2, if north (from which bearing angles are measured) at the receiver is parallel to north at the transmitter, then the desired line of position bearing is the exact reverse or *reciprocal* of the observed bearing. By reciprocal angle is meant the original angle plus 180 degrees.

At longer distances, however, there are complications in using airborne direction finders. The curved shape of the earth must be considered and some thought given to the precise meaning of "direction." Directions or bearings from a given point are usually specified by the angle measured, generally clockwise, from the meridian or true north line at that point. This is because north (or south), as determined roughly by a compass, ultimately by celestial observations, is the only natural reference direction available independently everywhere on the earth. Unfortunately, as Fig. 3 attempts to show, true north is not a constant reference direction always parallel to meridians or north lines at other places on the earth. Geographical north is a horizontal direction line on the earth's surface, and these lines converge toward the poles of the earth. Especially at high latitudes, the north line of an observer distant in longitude from a transmitter, is not parallel to the north line at the transmitter. The convergence, or amount of deviation from parallelism of the two meridians,

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[1] "Practical Air Navigation," CAA Bulletin 24, 1945 Edition; p. 229.

[2] Stewart and Pierce, "Marine and Air Navigation," Ginn and Company, New York (1944); p. 237.

could be calculated and allowed for—provided one knew the precise location of both places. But in a navigational problem, by definition, the observer is ignorant of his own position; that is just what he wishes to find out.

The direction-finding observer is thus in this peculiar position: He has measured the arriving direction, with respect to his own compass north, of the radio ray that reached him after traveling from the transmitter along a direct path, but that knowledge alone does not tell him the departing direction, relative to the transmitter's compass north, at which that same radio ray left the transmitter. The transmitter emits radio rays in all possible directions at once; to plot a line of position, the observer must be able to identify that one of them which reached him at the observed direction-finding angle.

One partial way out is open to the navigator if he knows his position approximately, as by dead reckoning. Then by visual estimation or calculation or by reference to special tables compiled for just that purpose, he can find out the convergence between the two meridians in question. The departing direction of the ray at the transmitter is then equal to the reciprocal of the direction-finder bearing of the same ray on arriving at the airplane, plus now the appropriate meridian convergence correction. This ray could then be plotted as a great circle bearing on a suitable map.

Apart, however, from the fact that plotting a great circle on a map is not always easy, there is an additional complication. The *great-circle path of the radio ray from transmitter to receiver* is not itself the *radio line of position* corresponding to the airborne direction-finder observation. Fig. 3 illustrates how a number of radio rays may leave a transmitter at different bearings, each one traveling along a different great-circle path, and yet arrive at a number of different points at the identical incoming or direction-finding angle. The curved, non-great-circular line drawn so as to join these points smoothly is the line of position that is characteristic of a direction-finding bearing measured on an airplane. To draw it accurately is a complicated process, involving also knowledge of the geometric properties of the particular map projection being used.[2,3]

At short or medium distances, where airborne direction finding is chiefly used, the various complications may be neglected partially or completely without serious error. Since this survey is concerned with navigation at all distances, at least the existence of the complications must be mentioned. If full understanding is still lacking, that fact may illustrate the point that navigation by direction-finding methods is only deceptively simple, at least at long distances. In any discussion of long-range navigational aids, one cannot escape the three-dimensional problems caused

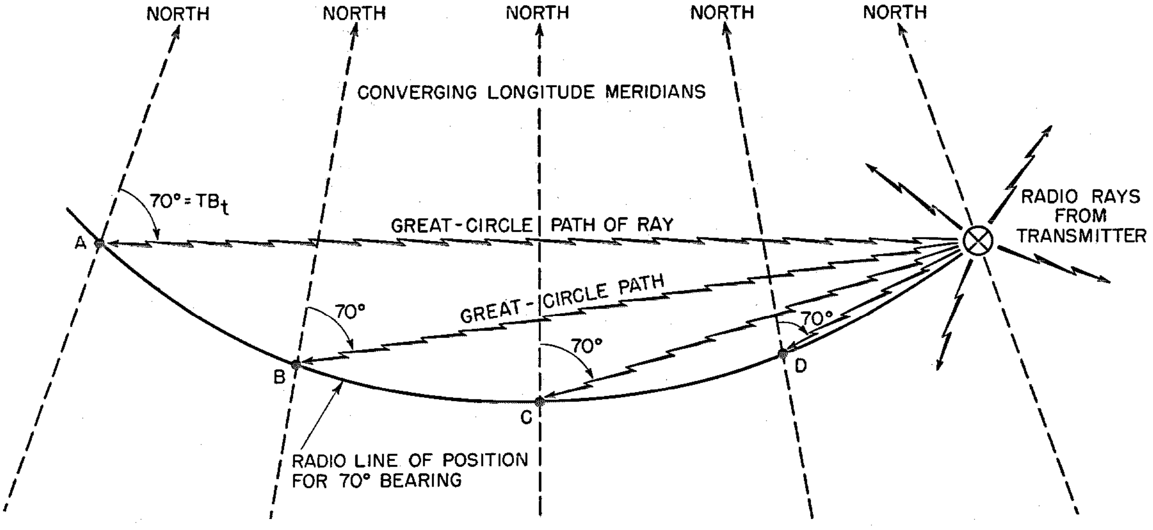
NORTH
NORTH
NORTH
NORTH
NORTH
CONVERGING LONGITUDE MERIDIANS
70° = TB$_t$
GREAT-CIRCLE PATH OF RAY
RADIO RAYS FROM TRANSMITTER
A
70°
GREAT-CIRCLE PATH
70°
70°
B
C
D
RADIO LINE OF POSITION FOR 70° BEARING

Fig. 3—Direction-finding complications at long distances. $TB_r = TB_t + 180$ degrees + meridian convergence correction.

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[3] P. 233 of reference 1.

by the sphericity of the earth. One may hope that four-dimensional space ships and navigational methods are never invented.

2.3 Ground Direction Finding

The situation is improved in some respects if the airplane transmits to two suitably located ground receiving stations. In this case, the true direction of the airplane, as "seen" from each of two known positions, is measured directly. That direction, laid out from the meridian at each ground station, is the radio line of position of the airplane. It is a great-circle arc and may be drawn as a straight line on a suitable map, viz.: on any projection for short distances, on a Lambert projection up to moderately long distances, and on a Gnomonic projection in the general case. The fix, as created by the intersection of two great-circle lines of position, might also be determined without graphical plotting by spherical trigonometry or by special devices.

Ground direction finding is frequently used for locating surface ships by reference to coastal receiving stations; it is sometimes, especially in emergencies, used for locating airplanes. For military purposes of spotting enemy ships, submarines, or airplanes, this method is fine since it requires no cooperation from the craft other than the emission of any type of radio signal. The "Huff-Duff," or high-frequency automatic direction finder, developed by Federal, is an example of this application, bearings being readable even on the shortest "squirts" of radio energy to which transmission was purposely limited to confound conventional, slow-reading direction-finding devices. But this is for special military purposes; it is applicable to ground observation of air traffic, a function important in itself but not, strictly speaking, a navigational aid. For intentional navigational use by the craft itself, ground direction finding and related methods (such as ground-based radar) suffer from certain fundamental drawbacks.

First, the distance range is more limited than in airborne direction finding because an airplane cannot carry a very powerful transmitter. Second, the airplane must first get in radio contact with a ground station to request its help; this and another suitably located ground station must take bearings on the airplane simultaneously, compare notes, plot the fix, and then radio the answer back to the airplane, which by this time may have moved. This procedure not only takes time, which is more serious with aircraft than with ships, but it also forces the navigator of the craft to depend on second and third parties, located at other places and less vitally concerned than is the navigator. This sort of thing is fundamentally objectionable in any navigational aid. Also, another return radio link is involved. Finally, ground direction finding is extremely "saturable"; that is, only one airplane at a time can use the ground stations' facilities. In summary, direction finding of the airborne or ground type is not a self-sufficient or independent navigational aid; it requires either compass indications or relaying of information from the ground.

2.4 Direction Finding Versus Range or Track Systems

Although suitably supplemented direction finding is ultimately capable of giving a fix, that ability is not the only item in navigation, or the one wanted most of the time. Especially on long overseas flights, the pilot is concerned not so much with where he may be exactly, as with the constant problem of how to steer so as to reach his destination most directly. Steering by only a compass or radio direction finder leads to troublesome off-course wanderings caused by variable wind effects, which mount up on long flights.

The pilot would like a radio navigational aid which gives him continuous track guidance or a sort of invisible radio beam to follow. He would like a *single* instrument to tell him continuously if he is on the prescribed flight path, and if not, whether he should steer left or right to return to that flight path. Such an instrument could also be arranged to guide the airplane along the flight path automatically, in conjunction with an automatic pilot. The essential thing is that each radio indication, entirely by itself, should correspond to a fixed line of position on the earth. Such systems, in contrast to directional receiving systems, are commonly referred to as radio range systems.

There may be some confusion in the use of this term, for one common meaning of range is *distance*, as in gunnery and radar. It is true that some radio navigational systems are of the pure distance type, giving circular lines of position, but, in most radio ranges, the term is used by way of analogy with the marine range principle, by which a pilot can guide his ship along the fixed straight-line path established by the visual alignment of two distant landmarks. In fact, in such radio ranges, the line of position is in effect delineated by two or more landmarks—antennas in this case—suitably aligned, the "sighting" being done by one of a variety of radio means.

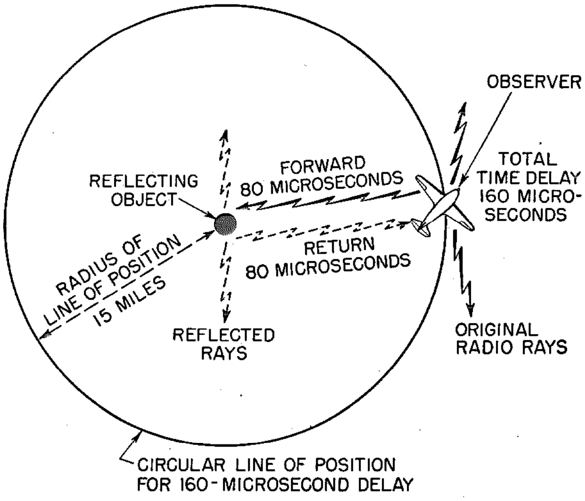
OBSERVER
TOTAL TIME DELAY 160 MICRO-SECONDS
FORWARD 80 MICROSECONDS
REFLECTING OBJECT
RETURN 80 MICROSECONDS
RADIUS OF LINE OF POSITION 15 MILES
REFLECTED RAYS
ORIGINAL RADIO RAYS
CIRCULAR LINE OF POSITION FOR 160-MICROSECOND DELAY

Fig. 4—Absolute-distance principle used in Shoran system.

Perhaps the most familiar and widespread example of a radio range or track guidance system is the four-course aural-type A–N installations operated all over the U.S.A. by the Civil Aeronautics Administration. A variety of other radio range schemes exist and give differently shaped lines of position. An ideal range sytem should not be limited to a specific number of beams,[4] but rather give omnidirectional service; that is, the pilot should have track guidance available completely around a ground station. Then at all points he could use intersecting beams or lines of position from two stations to get a fix whenever desired. Furthermore, the shape of the beams should be such as to offer convenient and direct flight paths between air terminals.

The three remaining basic types of radio navigational systems are each capable of giving range-type service or track guidance; each has a characteristic shape of line of position. Questions of omnidirectional service and other features of convenience depend on specific system details.

3. Type B: Absolute-Distance Systems

3.1 General Principle

The principle is that the distance between an airplane and a remote point is determined by measuring the time it takes a radio wave to make a round trip between the two points, one of them being at a known position on the earth.

3.2 Pulse Method (Shoran, Oboe)

In Fig. 4, an airplane emits a brief pulse of radio energy in all directions. Some of this radiation hits the known object and causes many reflected rays to be emitted in all directions from that point. One of these reflected rays travels back to the airplane, where it is received. This all takes a very short, but measurable, time. If the reflected pulse arrives back, say, 160 microseconds after the original pulse went out, the one-way passage must have taken 80 microseconds. Since radio waves travel at very nearly 186,000 miles per second, or 0.186 mile per microsecond, by simple arithmetic the distance between the airplane and the reflecting object must be close to 15 miles.

From this single observation only distance, not direction, is revealed. Therefore the airplane must be on a line of position which is a circle 15 miles in radius centered about the plotted position of the reflecting object. The procedure for drawing a circular line of position on a map would depend on the projection used; special charts could be drawn up in advance showing numerous lines of position corresponding to various time observations. By flying so as to keep a steady time-delay indication, the pilot could accurately fly along any desired circular flight path. This radio beam is a distance type of range, analogous to piloting by means of an optical (parallax method) or acoustic (echo method) range or distance finder on board a surface ship. To get a fix, the navigator would observe the time delay from a second reflecting object and note where the two circular lines of position intersect.

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[4] In radio range systems, the term "beam" is perhaps misleading but it is common; it really stands for the expression "observable radio line of position."

Shoran, developed by the Radio Corporation of America, and various types of distance-measuring equipments (DME) work essentially on the principles just described.

Alternatively, the primary transmissions and the time measurement could take place at a ground station, the airplane itself, or a responder-beacon on it, serving as the reflecting object. This is the principle of the British Oboe system. There enters the problem of relaying the distance or fix information, determined on the ground, back to the airplane via another radio link. Lines of position are still circles centered on the ground station. Service radius is still limited by the requirement of reflections (or aided responses) from the airplane. It might also be noted that circular flight paths are not especially useful, except for "orbiting" or "holding" maneuvers around airports, preparatory to landing. These methods, however, have proved very useful for military and other applications demanding great accuracy in pin-pointing a fix as there are very accurate timing methods available.

3.3 Timing Principles

Measurement of the microscopic time intervals involved is done by special electronic means, generally in conjunction with a cathode-ray tube whose screen performs the function of the dial of a watch. An electric current is caused to make an electron beam behave like the sweep-second hand of a watch. The position of the electronic hand or beam can be made visible on the cathode-ray-tube screen. The hand starts moving from zero each time the basic pulse ("main bang") goes out, and sweeps along the time graduations at a uniform rate. Some time during this sweep, the echo pulse arrives and illuminates the hand at that instant by causing a "pip" of light; the position of this pip against the scale of uniform time graduations indicates the time interval, as when using a watch. In some cases, the electronic watch hand sweeps around in a circle continuously like an *ordinary* watch hand (circular time-base sweep); in most cases the electronic hand moves laterally or radially outward in a straight line, and at the end must be very quickly reset to zero as when one rapidly double-clicks a *stop* watch to start the process over again (linear sawtooth time-base sweep). In each case, the hand must travel at a uniform speed, and care must be taken to have it start from zero each time the main pulse goes out. The electronic hand moves too rapidly to allow one to follow its motion directly, therefore it is kept dark and its momentary position is made visible only when pulses occur. Each single pip or visible pulse is too brief to notice, as in the case of a single frame of a moving-picture film; and as in moving pictures, repetition, together with persistence of vision, solves this problem. Two ways of indicating such time intervals are shown in Fig. 5.

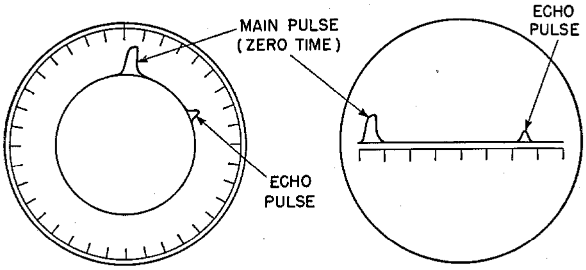

Fig. 5—Principles of two common ways of displaying on a cathode-ray tube the time interval between pulses. Graduations may indicate time in microseconds or the corresponding distance in miles.

In some cases, controls are provided to move the echo pip into apparent coincidence with the main pulse pip, and the time is given by the position of the control. In certain applications, such as radio altimeters and distance indicators, the electronic circuits may be arranged to produce a direct-reading mechanical meter-type indication, generally not so accurate. In all cases, the ultimate or theoretical limit of accuracy with which the exact time of occurrence of a pulse can be determined depends on how short the pulse can be made. As the minimum length of a pulse is inherently limited by the bandwidth of the radio channel, increases in ultimate timing accuracy for radio distance measurements require wide channels which are practically available only in the higher-frequency region of the radio spectrum.

3.4 RESPONDER-BEACONS (IFF)

There must be some provision for distinguishing those echoes that are produced by the desired reflecting object at the known location from the mass of echoes that are produced by reflecting objects located at various other points in the path of the primary radiation. Where airplane interrogators are used, the problem may be solved by installing responder-beacons at the desired ground locations. These beacons consist of a receiver-transmitter combination and are also called transponders and racons. Military code names, often of a fanciful character, have been applied to them. Reception of each interrogation signal or pulse automatically triggers the transmitter into radiating an answering signal. This answering signal (active response or artificial echo) is stronger than a passive or natural echo, so that the range of service is increased. For positive identification of the different responder-beacons, the answering signals may be coded distinctively or may occur on different radio channels.

In systems where the primary transmitter or challenger is located on the ground, responder-beacons on the airplane are valuable for increasing the range, as natural echoes from airplanes are quite weak, and for identification purposes. Identification-of-friend-or-foe (IFF) systems, used in wartime to identify airplanes spotted as pips on radar displays, worked on these principles; friendly airplanes were provided with responders which coded their answering pulses in a prearranged manner.

A single responder-beacon may provide automatic answering service for a number of different interrogators at the same time, but there is a limit. One factor determining the saturability of the system is the power capability of the transmitter portion of the responder; if the energy per response pulse (which affects the range) is not to fall off, the total number of pulses per second that the transmitter may radiate is limited by its rated (average) power output. A second factor holds true also for systems using natural echoes; there is a limit to the number of different sets of interrogation or response signals that can exist on a given radio channel without likelihood of interference, even though each set may be composed of brief and widely separated pulses.

3.5 PHASE AND FREQUENCY-MODULATION METHODS

Absolute distance may be determined in other ways than by using brief pulses; but in all cases transmitted and reflected energy must be used and the observation is fundamentally, if not directly, a measurement of *time.*

The transmitter may emit continuous-wave radio energy. In this case, the difference between the instantaneous phase of the departing wave and the phase of the wave arriving back after reflection is measured. This phase difference depends on the total number of wavelengths and fractions thereof included in the total round-trip distance; this in turn depends on the total distance and the known wavelength or frequency of the radio signals. In all cases of wave motion, the instantaneous phase changes uniformly by 360 degrees per wavelength as one travels along the path of the wave. The phase at any one point also varies with time, according to the radio-frequency rate; but at all times the *difference* in phase between any two points is constant, depending only on the distance between the points in terms of wavelength. Thus measurement of phase differences may be a convenient and sensitive way of determining distance. This principle is used in one form of the Raydist system.

A related method is to vary the radio frequency of the outgoing radio waves, usually at a sawtooth rate, and to compare the instantaneous difference in frequency between outgoing waves and reflected incoming waves. This difference depends on the time of travel of the waves and is measured by some type of differential frequency meter. This principle is used in some forms of radio altimeters or terrain-clearance indicators (TCI).

3.6 RADAR PRINCIPLE (PPI DISPLAY)

Radar is a combination of types but is logically mentioned at this point because one part of it is always an absolute-distance system, generally of the pulse type. In radar, defined as a system for "radio direction and ranging," a complete fix is obtainable by *simultaneous* determination of a circular line of position from the absolute-distance system and a radial straight line of position from the directional system, both based on the *same* landmark.

The principle is illustrated in Fig. 6. The bearing or directional line of position is obtained by the use of directional antennas for receiving or for transmitting. Actually, most radar systems use the same antenna for both emitting the original radio waves and receiving the reflected waves. Brief pulses are transmitted along a sharply defined narrow beam[5] instead of in all directions. The direction of this beam is gradually changed by physical rotation of the antenna, which is therefore said to scan in all directions. The echos are displayed pictorially as spots of light on a cathode-ray-tube screen. Radial distance of the spot of light corresponds to the time delay or actual distance of the reflecting object. The direction at which the spot appears corresponds to the actual bearing of the antenna at the instant of echo reception. All distances and directions are with respect to the position of the antenna installation, which corresponds to the center of the display. This type of display is a polar-coordinate plot and is referred to as a plan-position indicator (PPI).

If the radar set is on the airplane, bearings are relative to the axis of the airplane and require a compass correction for absolute or true orientation. On a ground radar installation, the true bearing of the airplane from the landmark is indicated directly. This is satisfactory for ground observation or traffic-control purposes, but if the ground radar is to constitute a navigational aid, both the distance and the directional information must be relayed back to the airplane via another radio link.

Radar antennas, directional and capable of rotation, add considerable weight and drag to airplanes. In any case, they require the use of rather high-frequency radio waves to avoid extreme size. This, in turn, limits the service range approximately to line-of-sight or horizon distances, even for ground radar. Thus, while absolute distance and radar systems are spectacularly useful for military purposes and appear promising for ground control of air traffic, as navigational aids they seem to be restricted to short- and medium-distance applications.

For military applications, normal (primary or unassisted) radar is particularly effective because no cooperation from the airplane is necessary. For civil purposes, the performance of ground radar, in respect to clarity of presentation and range of service, can be improved by installing responder-beacons in aircraft. The artificial responses are stronger than natural echoes and, if they are on a different radio channel, the radar receiver can reject all natural echoes and ground clutter to produce a clear display. The response, being artificially produced on the airplane, can be coded to convey information such as the altitude of the airplane and its identification. Thus, beacon (secondary or assisted) radar will doubtless find extensive use in civil air-traffic-control systems.

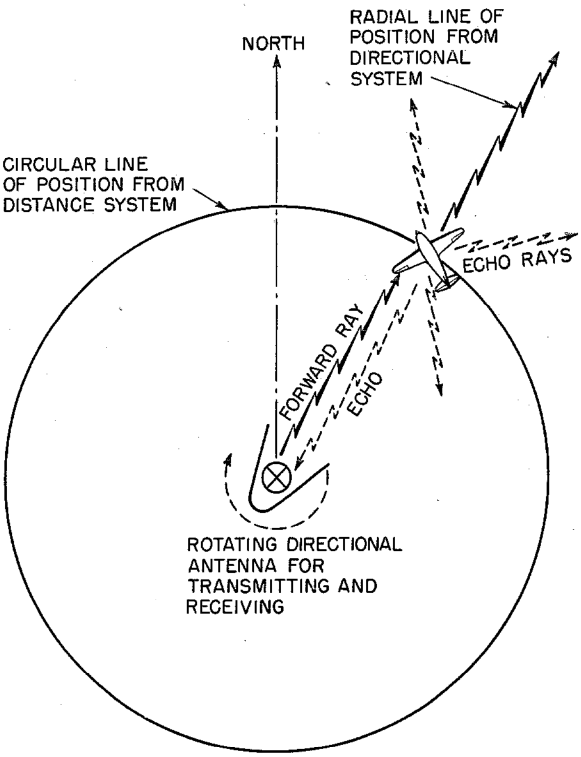
NORTH
RADIAL LINE OF POSITION FROM DIRECTIONAL SYSTEM
CIRCULAR LINE OF POSITION FROM DISTANCE SYSTEM
ECHO RAYS
FORWARD RAY
ECHO
ROTATING DIRECTIONAL ANTENNA FOR TRANSMITTING AND RECEIVING

Fig. 6—Radar principle using plan-position type of indication.

4. *Type C: Differential-Distance Systems*

4.1 General Principle

In this type of system the *difference* in distance between an airplane and two different ground stations is determined, but again making use of the known speed of travel of radio waves.

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[5] In this case, the word "beam" for radio line of position is used with justification.

4.2 Pulse Method (Loran, Gee)

Fig. 7 illustrates the essential principle. Two ground transmitters, *A* and *B*, each emit brief pulses of radio energy at exactly the same time. To reach an airplane located at *X*, the signals travel along different paths. If the lengths of the two paths are different, one pulse is received later than that from the other ground station. This time difference or delay is a measure of the difference in path length, or *BX* minus *AX*.

In Fig. 7, the 300-microsecond delay observed at the airplane reveals only that distance *BX* minus *AX*, or distance *BY* minus *AY*, and so on, equals about 55 miles. Many different points may correspond to that same time or distance difference. The exact line of position depends on the time difference and on the location of the two transmitters, but in any case is one of a family of hyperbolas (by the geometric definition of such curves) on a plane surface. The line of position of no time difference, or of equal path length, to express it more generally, is clearly the perpendicular bisector of the line joining the two ground stations.

This is the essential principle of Loran, developed by the Radiation Laboratories at Massachusetts Institute of Technology; Gee, developed in Britain; and a number of related systems, all sometimes referred to as hyperbolic systems. Organizations that have built or are building Loran equipment in the U.S.A. include R.C.A., General Electric, Sperry, Federal, and others. One great advantage of differential- over absolute-distance systems is that there is no need for a return signal from the airplane. This increases the practical service range. Moreover, the system is not saturable for any number of airplanes can use the signals simultaneously. The observations are made directly on the airplane.

Other basic characteristics include the following: Simple nondirectional antennas are used. Stations *A* and *B* must be separated by a considerable distance, up to several hundred miles, or the time differences become too small to measure accurately. Station *B* must constantly receive signals from station *A* for control purposes for the times of emission of signals from both stations must be accurately synchronized. To cover a given area with one family of lines of position, two suitably related ground stations are required and an airplane must be within reception distance of both stations.

A fix is established by observing the time delay with respect to another pair of ground stations within range and noting the intersection of the two hyperbolic lines of position. The two *pairs* of stations may have one station in common. Special charts are required to facilitate the interpretation of time delays into lines of position on a map. These charts have ordinary latitude and longitude lines plus the hyperbolic lines corresponding to various time delays from a number of pairs of stations in the area, with allowance for the curvature of the earth and the properties of the map projection all figured in advance.

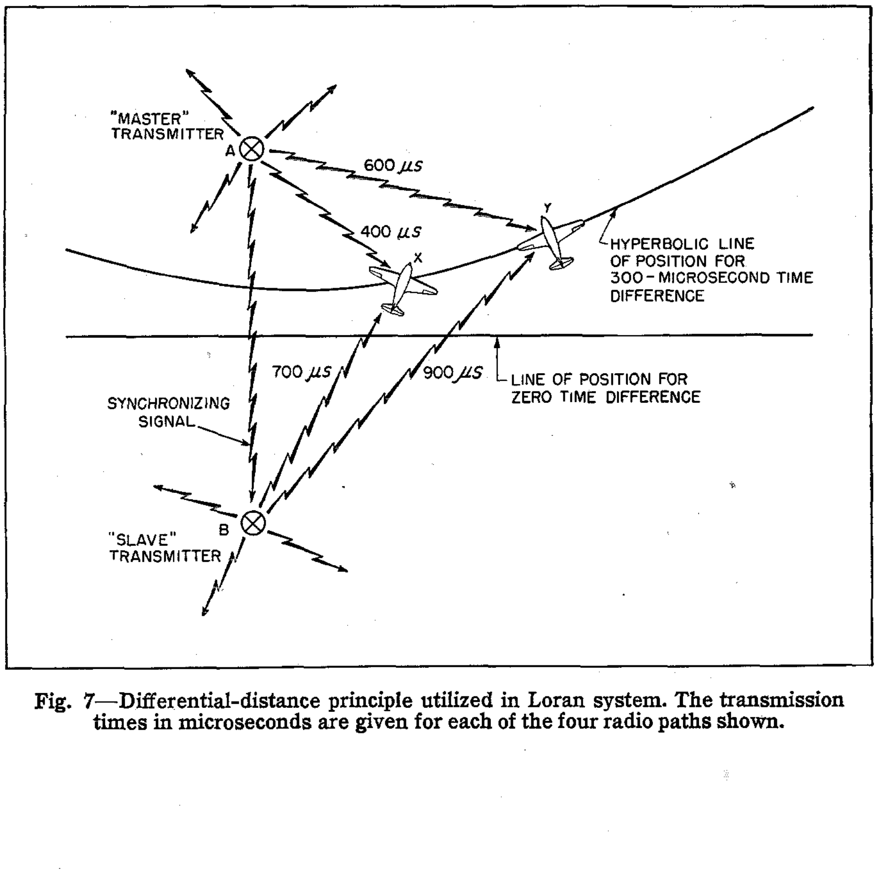
"MASTER" TRANSMITTER
A
600 μs
400 μs
Y
X
HYPERBOLIC LINE OF POSITION FOR 300-MICROSECOND TIME DIFFERENCE
700 μs
900 μs
LINE OF POSITION FOR ZERO TIME DIFFERENCE
SYNCHRONIZING SIGNAL
B
"SLAVE" TRANSMITTER

Fig. 7—Differential-distance principle utilized in Loran system. The transmission times in microseconds are given for each of the four radio paths shown.

Although Loran requires no rotating or directional antennas and is thus not restricted to very-high-frequency line-of-sight applications, the fact that brief pulses are used places some restrictions on the operating wavelength and bandwidth. The wavelengths at which Loran is usually operated allow service up to about 700 miles by day and about 1400 miles by night, but, unfortunately, the transmissions are subject to the ionospheric disturbances that plague all high-frequency radio signals.

By flying so as to keep a constant time delay, the pilot can guide his airplane accurately along any line of position as a track. Hyperbolic lines, however, are not generally suitable as direct flight paths. The central or zero-delay line of position is an exception, being a straight line or great-circle path. But for this line to pass through a given air terminal requires siting ground stations at two additional locations. This may not always be practicable in remote regions or at islands in midocean; and, in any case, none of the other lines of position "lead home." Indication of time delay is not automatic but requires skillful and time-consuming observations and manipulations of cathode-ray-tube devices. This, however, is an engineering detail which may no doubt be improved; it has certain advantages in fact in that it provides a visual indication of the quality of the signals for monitoring purposes.

With all these drawbacks, differential-distance systems have been put to extensive use for short and medium-long ranges, especially in military applications, because of their great accuracy and reliability up to a certain variable limiting range. Whether reliable world-wide coverage can be obtained, either by improving the dependable distance range or by finding sufficient pairs of suitable land sites, is a question that has not yet been answered.

4.3 Phase Method (Decca, Raydist)

Differential-distance systems, like absolute-distance systems, may employ continuous waves instead of pulses, with phase comparison instead of direct time-delay measurements. This is the principle of one form of Raydist, an accurate system suitable for surveying, and of the Decca system developed in Britain by the company of that name. In the latter, master and slave stations emit pure continuous-wave radiations accurately synchronized in radio-frequency phase. At any distant point, the difference in phase between the two waves received on an airplane depends on the difference between the number of wavelengths, and fractions thereof, in each path. The lines of position are still hyperbolas and the system has most of the other basic characteristics of Loran, except of course that there are no inherent limitations on the operating frequency or bandwidth. Other practical complications do arise. These involve the question of ambiguities and the special relations between the radio channels of a pair of stations.

Differential-distance systems of either the pulse or the phase type may alternatively work on the principle of having two separate ground stations receive signals from the airplane. Lines of position are still hyperbolas. A link is now required for comparing the time or phase of reception at the two ground stations; an additional radio link is needed to relay the information back to the airplane. Range is limited by the necessity for an airborne transmitter; and, like ground direction finding, the system is saturable. While perhaps useful for some special military applications, as in Micro-H, this method is clearly unsuitable for a long-range self-sufficient navigational aid.

5. *Type D: Directional Transmitting Systems*

5.1 General Principle

Here the fundamental principle is that the transmitting antenna system has directional characteristics; that is, the radio signals transmitted in various directions from the station are different in some measurable respect.

If the transmitter is located at a ground station, the radio lines of position are fixed straight lines or great circles radiating from that point. If airborne, type-D transmitters are generally limited to radar applications where antennas may be small; range of service is then short so that meridian convergence is negligible. The lines of position, however, are relative to the fore-and-aft axis of the airplanes as in airborne direction finding (see Fig. 2); but just such relative indications are useful in locating nearby obstacles for anticollision service.

Antennas are given directional properties by making them large compared to the wavelength or by using arrays of antennas suitably spaced and phased. The principle is the same as in any wave-interference phenomenon, as occurs also in optics and acoustics, in which the waves from the individual sources annul or reinforce each other depending on how they combine in phase along different directions.

Radar antennas are a special case in which, because of the high frequencies used, the antenna "dish" or array can be very large compared to a wavelength and yet be small or moderate in physical size. Such antennas can produce practically one single sharp beam, and the radio line of position is simply the direction in which the antenna is pointing. Being small, the antenna can be physically moved around so as to point in any desired direction. Service range, however, is limited to radio line-of-sight.

At lower frequencies, directional effects are obtained by the use of two or more antennas separated from a quarter to several wavelengths. Actual size is large and practically precludes physical rotation. Sharp beams are not produced by such systems. At these frequencies, common examples of directional radiation patterns or graphs, which indicate the relative signal strength in various directions, are shapes like the figure-of-eight and the cardioid.

This variation of signal *strength* cannot, by itself, be used immediately to indicate direction. This is because the actual signal strength existing at a given point depends also on transmitted power, distance, and propagation conditions, which factors cannot be expected to remain constant. They must be made to cancel out in some manner so that only the effect of the directional properties of the transmitting antennas on the signals remains. A radiation pattern, in fact, indicates only the relative strength of signals in different directions if all other factors are assumed to be constant. Fortunately these other factors are constant at a given point for short time intervals at least.

Therefore in these systems, the transmitter always emits at least two types of signals, corresponding to two different directional patterns, so that regardless of the actual strengths of the two signals at a given time and distance, their *relation* to each other is constant and depends only on the bearing of the observer. Both signals may vary for one reason or another—even receivers vary in sensitivity—but all these variations affect both signals in the same proportion provided the two types of signals are emitted simultaneously or in very rapid alternation.

At the frequencies in question, it is impracticable to produce different directional patterns by physical movement of the antennas. Instead, they are produced by electrical means; that is, by switching to different pairs of antennas or by changing the electrical excitation (current strength, phasing, or both) of a given set of antennas. Thus, with physical movement limited to small electrical control devices, the lobes of the directional patterns from a fixed antenna installation may be rotated, shifted, reversed, or interchanged, and so on, and at a fast rate. As a result, along each direction two or more signals of different strength, or other measurable characteristics, are repeatedly provided, which the observer compares and from which comparison the true radio line of position is determined.

The two or more types of signals, corresponding to different directional patterns, must be distinguishable at the receiver so as to permit comparisons to be made. For this purpose, the radio waves corresponding to each directional pattern are given some characteristic modulation or other identifying feature, such as tone modulations, dot-and-dash code patterns, or certain sequences or durations of emission. The method of separating and comparing these signals and of indicating their relation depends on specific system details. The comparisons are not always directly in terms of relative signal strength, but may be in terms of other characteristics such as modulation patterns (phase, code sequence) which, however, depend on the directional strength patterns. Also, usable signal relations may exist only along a limited number of directions as in single landing beams and two- or four-course ranges, or at all bearings as in the omnidirectional ranges (ODR).

5.2 Limited-Coverage Ranges (A-N Range, Landing Beam)

The four-course A-N ranges, though considered old-fashioned, are the most extensive radio navigational aids at present in the U.S.A., and may be considered as a sort of prototype of type-D systems. The indications are aural and the airplane needs no special equipment other than a communications radiotelephone receiver and headset; but there are disadvantages in that the indications in such methods are either inexact or time-consuming.

In the A-N range, two pairs of antennas are used, each located at opposite corners of a square. Spaced from a quarter to a half wavelength apart, they are excited so as to produce two figure-of-eight patterns at right angles. A dot-dash *A* code character is emitted on one pattern and a dash-dot *N* character on the other pattern. The two figure-of-eight patterns intersect along four bearings or beams where, for equal power in each radiation, both signals are equally strong. Actually a continuous tone is detected since the dot-dash alternations are suitably interleaved; this is the aural "on-course" indication of the radio line of position. Along other directions either the *A* or the *N* code stands out, and indicates on which side of the beam the airplane may be. No means is provided to make exact quantitative comparisons of strength. Thus, only four useful fixed beams result as will be seen from Fig. 8. Navigation along other routes is not provided for; fixes by means of intersecting beams from other stations are only occasionally possible. The four beams sound alike, so if the pilot is completely lost, considerable ingenuity and time are required to judge which of the four beams he is on—if at all possible.

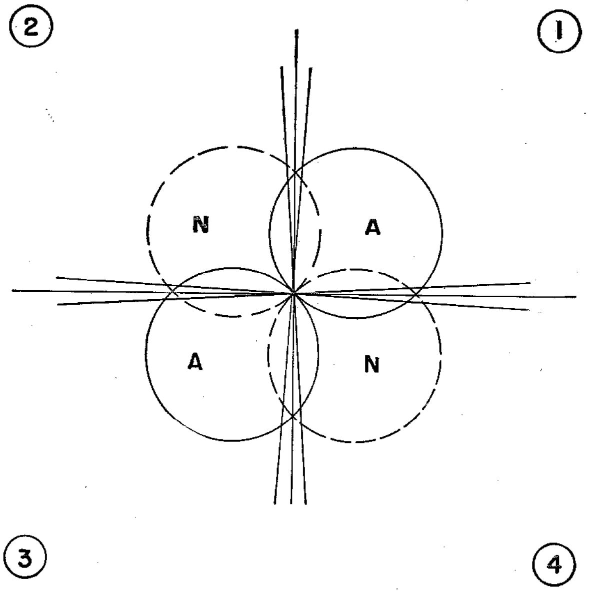

Fig. 8—Principle of the A-N radio range system showing the two figure-of-eight patterns alternately radiated and the four equisignal sectors or beams.

The equisignal intersecting-lobe or lobe-switching principle is a fundamental one, and is often applied. Landing systems and other two- or four-course range systems use it, even though the indication of equality or balance may be presented visually on a meter. In landing systems a single, but very accurate, beam is all that is desired. For this purpose, larger arrays, in terms of wavelength, are used to produce sharp lobes, only one pair of which are used. The alternating radiations have different audio-frequency modulations which can be separated by filters in the receiver and applied to a zero-center meter to pull the needle in opposing directions. The principle is even used in some radar receiving antennas to increase the directional sharpness, for radar beams have some width; two intersecting lobe patterns, at a slight angle to each other, are produced electrically from one antenna, which is accurately lined up by using some sensitive balance indicator.

5.3 INTERMEDIATE-COVERAGE RANGES (CONSOL, SONNE)

Greater coverage of a given area with useful beams or radio lines of position can be provided if directions, in general, are quantitatively indicated in terms of the measurable ratio of strength between the alternating signals. Operation is not restricted then to equisignal directions. Coverage is still not complete, because some portions of the lobes are not useful; at the nulls or minima, signals are too weak to be heard; at the maxima, the variation of strength with direction is too gradual to permit accurate determinations. If, however, the available lobes from a given set of antennas are rotated or shifted enough so as to cover all sectors at one time or another with useful portions of lobes, continuous coverage is possible. Complete omnidirectional service is still not possible by any schemes using one pair of antennas; the two angular sectors along the line joining the antennas turn out to be "dead zones" in this respect.

The principles just described are used in the British Consol system developed from the German Sonne system. Equipment for an American form of the system has been built by Federal. This system is also aural, requiring only a radiotelephone receiver and headset; great accuracy is possible because the lobes are narrow; and continuous coverage of wide, but not unlimited, areas is provided. The readings, however, take considerable time, special reference charts are required, and there are many ambiguities which depend on the number of lobes.

Two or three in-line antennas are used, spaced up to as many as six wavelengths apart, resulting in many narrow lobes. Successive, intersecting pairs of lobes are produced by rapidly alternating reversals of current. For identification purposes the complementary lobe patterns are radiated in a sequence of alternating dots and dashes. At a slower rate, and with the rapid lobe switching meanwhile continuing, the lobes are gradually swept around in bearing by means of gradual, uniform changes in phasing. The result is that an observer at a given bearing hears a peculiar pattern of dots and dashes. The apparent coding sequence that is heard depends on the observer's bearing. He counts the number of dots and the number of dashes heard during some specified time interval, consults a special calibration chart, and reads off his true bearing.

The nature of the radio emission is such that the system may operate on any radio frequency, even very low; and very narrow bandwidths may be used. The ambiguities and the lengthy aural presentation, together with the lack of complete coverage, are the chief drawbacks.

5.4 OMNIDIRECTIONAL RANGES (NAVAGLOBE, CAA OMNIRANGE)

For omnidirectional service, so that beams are available through all 360 degrees, and so that fixes by crossed beams are everywhere possible, more antennas aligned in different directions are needed. To avoid the need for an excessive number of antennas, the previously described principles of lobe sweeping or of non-equisignal measurements may be used at the same time.

For very long distances, the straight lines of position of ground-based directional transmitting systems are desirable, but no entirely suitable systems of this type have been installed. The main problem is devising a system which is propagationally (frequency, bandwidth, etc.) suitable for very long distances and which is also omnidirectional, direct-reading, and free of troublesome ambiguities.

One interesting example of a system for that application, and which serves to illustrate the general principles of directional transmitting methods, is the Navaglobe system,[6, 7] now being developed. This was proposed by Federal at the national radio navigation conference in Washington in February, 1946.

Three adjacent antennas are used for the ground station installation. Three is the minimum theoretical number of antennas capable of giving omnidirectional service. The antennas are spaced triangularly, as shown in Fig. 9, about 0.4 wavelength apart. In turn, each pair of antennas is equally excited so that in effect three signals are radiated in rapid succession, over and over again. The relative strength of each signal depends on the direction of the receiver from the array; this is indicated graphically by the

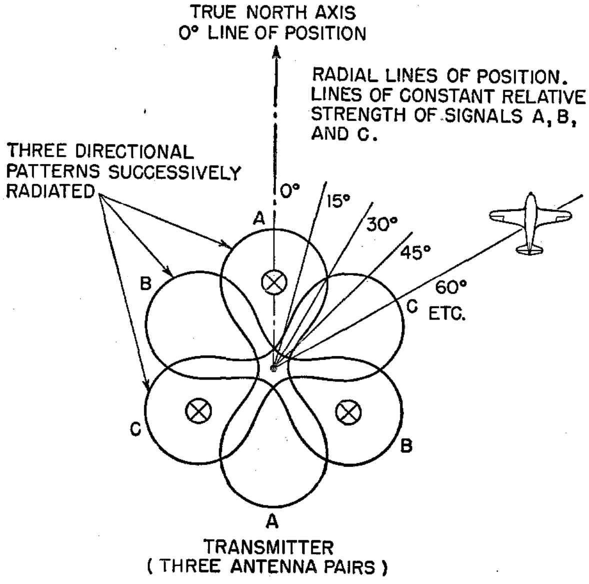
TRUE NORTH AXIS
0° LINE OF POSITION
RADIAL LINES OF POSITION. LINES OF CONSTANT RELATIVE STRENGTH OF SIGNALS A, B, AND C.
THREE DIRECTIONAL PATTERNS SUCCESSIVELY RADIATED
0°
15°
30°
45°
60°
ETC.
A
B
C
C
B
A
TRANSMITTER
(THREE ANTENNA PAIRS)

Fig. 9—Directional transmitting principle on which the Navaglobe system is based.

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[6] H. Busignies, P. R. Adams, and R. I. Colin, "Aerial Navigation and Traffic Control with Navaglobe, Navar, Navaglide, and Navascreen," *Electrical Communication*, v. 23, pp. 113–143; June, 1946.

[7] P. R. Adams and R. I. Colin, "Navaglobe Long-Range Air Navigation System," *Proceedings of the National Electronics Conference*, October 1946.

radiation patterns or relative signal-strength curves shown. Along each straight radial line, the *relative* strengths of the three signals received in rapid succession are constant; at greater distances all three signals are weaker, but in the same ratio. The receiving system measures this relation between signal strengths and translates it automatically into the setting of a pointer around a dial calibrated directly in degrees of azimuth or true bearing.

In common with all ground-based directional transmitting systems, and in contrast with both ground or airborne direction-finding systems, the indicated bearing is directly the bearing of the observer from the ground station, and is at the same time observed directly on the airplane. The radio line of position is that bearing laid off as a great-circle bearing from the meridian of the ground station. This is easily plotted on the Lambert projection, commonly used in government air navigational maps, as on that projection great circles may be represented by straight lines for even moderately long distances. In any case, a straight line may be drawn on the Gnomonic projection, which is also furnished by the government for air navigational purposes. For a fix, the bearing from another ground station is observed and the intersection of the two lines of position is noted. There are also devices which quickly give the latitude and longitude of a fix corresponding to two great-circle bearings from known points without need for maps or calculations. In short, ground-based type-D lines of position and fixes are as simple and straightforward as in ground-based direction-finding, yet they arise from ground transmitters and airplane observations instead of the reverse.

Furthermore, because the lines of position are great circles, they are useful direct flight paths or radio beams for traffic to follow. They all lead to or from the ground station; and if ground stations were located at or near the main terminals of long-distance air routes, the lines of position would be direct fixed homing paths.

In addition, as Navaglobe is an omnidirectional system, the pilot could select any desired radial direction as a flight path; and could use crossed lines of position for a complete fix at any place as long as he is within range of two stations. Moreover, since the Navaglobe indicator is a direct-reading pointer type, the pilot could easily "keep on the beam" by setting an index mark along the dial at any desired bearing and steering so as to keep the pointer aligned with the index. A differential left-right meter might be added to give him corrective steering directions still more conveniently, or to guide an automatic pilot.

The nature of the system is such that the operating wavelength may be of any value, even very long. This is because the emissions may be very slowly modulated continuous waves with their attendant extremely narrow bandwidths.

The Civil Aeronautics Administration has developed an omnidirectional visual radio range system, which is intended eventually to replace the four-course aural A-N ranges for medium-distance overland navigation. This new system uses higher radio frequencies than does the old aural system, partly because its modulation features appear to require it. The antenna array, incidentally, is smaller.

This omnidirectional range uses five antennas, four being spaced at the corners of a square, and one at the center. The effect of almost uniform circular rotation of lobes is obtained electrically by the rotation of small control devices or goniometers instead of by movement of the antennas. Suitable excitation of each pair of antennas at diagonally opposite corners of the square, in conjunction with the central antenna, produces a cardioid or heart-shaped directional pattern. By a scheme of modulation which fades the four cardioid patterns in and out properly at a slow audio-frequency rate, the effect is similar to that of a cardioid directional pattern rotating uniformly. The resulting signal heard at a distance is a wave which has a corresponding audio-frequency modulation imposed on it, but the phase of this modulation depends directly on the bearing of the observer. An additional audio-frequency modulation is imposed on the radiation through the central antenna, at the same rate; but this signal is nondirectional and has the same phase at all bearings. This serves as a reference signal.

The receiver on the airplane has automatic means for separating the two modulations and indicating the phase difference directly on a meter, in terms of true bearing. There are no ambiguities since the antenna produces a cardioid, or single-lobed pattern. One degree of phase difference corresponds, almost exactly, to one degree of bearing change. As there are no large moving parts in the transmitting system, the rotation rate can be fairly high and the reading time of the indicator correspondingly short.

The system has all the features of observational convenience described under Navaglobe; one basic difference is that its system of modulation is more complicated and requires, in present form, wide bandwidths and high frequencies. Thus, while it is propagationally suitable for medium distances, its suitability for very long ranges appears unsettled; development work, however, is being done on a low-frequency design of this system.

6. Ambiguities

6.1 General Nature

An ambiguity exists when a single navigational observation corresponds to two or more lines of position. Thus two observations may result in a fourfold or greater ambiguity of the fix. Ambiguities exist even in celestial navigation, as in the most common method (Sumner lines of position by means of zenith distances) wherein two circles of position intersect at two points. It might almost be said, in general, that all fixes are subject to an ambiguity caused by the fact that so-called straight lines of position are really great circles on the earth, and two great circles intersect at diametrically opposite points on the earth. A navigator, however, is not generally lost to the extent that he is ignorant of which hemisphere he is in.

It is possible, on the other hand, for the alternate lines of position or fixes to be so related as to cause troublesome ambiguity even to good navigators; this may be a major factor when appraising some radio navigational systems. Characteristic types of ambiguities occur in each type of system.

6.2 Directional Systems (Types A and D)

In these systems, ambiguities depend on how many lobes there are in the directional patterns. The Navaglobe pattern, for example (Fig. 9), has two diametrically opposite lobes per antenna and there is an ambiguity in indicated *bearing* of exactly 180 degrees. Simple loop-type direction-finding systems have the same ambiguity, for the same reason. In direction-finding systems, however, it is possible to "resolve" the ambiguity by the use of an additional omnidirectional or sense antenna, properly connected, which produces a single-lobed directional pattern, the familiar cardioid.

In either case, however, there is no ambiguity in *line of position*. To illustrate, a line of position of bearing 60 degrees is the exact backward prolongation of the line of position of bearing 60 degrees plus 180 degrees or 240 degrees. Both bearings result in the same straight line. Thus two bearings, each with a possible 180-degree ambiguity, can result in no more than two straight lines; and two straight lines can intersect at only one point. Therefore, there is also no ambiguity in the *fix*, except for the antipodal ambiguity caused by the fact that the lines are actually great circles.

Other systems of type D may have ambiguities in bearings by amounts other than 180 degrees, in which case there are necessarily ambiguities in lines of position and fixes. This, in general, results from radiation patterns with many lobes, close together. Such systems, on the other hand, are inherently capable of better accuracy; that is, there may be a greater, more accurately measurable difference between the radio indications corresponding to lines of position that are close together. The reason for this is that with many lobes, the lobes are narrower or sharper, the signal strength varying more rapidly for a given change in direction. Wider spacing of the individual antennas in terms of wavelengths, or increased number of antennas, causes this situation. In any case, it is the number of lobes that are physically indistinguishable from each other by the means provided in the system, which determines the number of ambiguities.

In the A-N type aural range, for example, there are four lobes, created by the two intersecting figure-of-eight patterns. The pilot can only distinguish, directly, the fact that he is on one of four possible equisignal intersections or radio beams. If he is completely lost and wishes to determine just which one he is on, rather involved mental (and aerial) gymnastics are required,[8] and are then not foolproof.

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[8] Page 207 of reference 2.

The Consol system uses wider spacing of antennas, up to six wavelengths, so that there are many narrow lobes. Accuracy is greater, but the line of position corresponding to a given observation is ambiguous to a manifold degree, for it may occupy similar positions along some of the other lobes.

The Civil Aeronautics Administration omnidirectional range is unambiguous because the antenna pattern is essentially a broad but single-lobed cardioid figure, as in direction-finding systems using a loop plus a sense antenna.

In directional transmitting systems of the radar type, great accuracy can be obtained without ambiguity by using antennas very large compared to the operating wavelength, so that a single very narrow lobe is produced. The range is limited to radio line-of-sight distances; and since only one lobe is produced, its movement to cover all areas must result from physical rotation of the antenna. In practice, this physical rotation cannot be very fast; not nearly so fast as with the purely electrical scanning described under longer-wave type-D omnidirectional systems. Also, increased radar-beam sharpness brings in other complications, involving practical limitations on pulse-repetition rate, transmitter power, and scanning rate.

If a set of observations corresponds to several possible fixes, the ambiguity may be resolved if, by some alternative means, one can determine his position with sufficient accuracy to judge that all but one of the fix locations is out of the question. The alternative means may be dead reckoning, rough celestial observations, or alternative radio aids. Or one might have avoided the ambiguity by having made observations continuously from some known point, and having kept track of them so as to be sure that one has not "slipped a whole cog" (lobe) en route. But if any such procedures are absolutely necessary, the navigational aid, whatever its other advantages may be, cannot be called an independent or self-sufficient system, one that requires no previous knowledge of position.

6.3 Pulse-Distance Systems (Types B and C)

In pulse-type absolute-distance or differential-distance systems, there is also a connection between the limiting accuracy of the system and the existence of ambiguities. The accuracy with which the time interval between two events can be measured depends on the speed of the time sweep. Though the time intervals are microscopic and the measurements are performed electronically on a cathode-ray tube, the principle is entirely analogous to using a sweep second hand of a stop watch.

A fast time sweep "stretches out the time," and is analogous to measuring the time between two swings of a pendulum by means of a watch hand that turns at a rate of 10 times per second. A slow time sweep "compresses the time" and is analogous to measuring the same time interval with a watch hand that creeps around the dial at a rate of once per hour. In the latter case, only a very rough measurement could be made of a brief time interval.

In pulsed radio systems, the basic pulses occur repeatedly at some fixed pulse-recurrence frequency (PRF); the time sweep or electronic stop-watch hand is reset to zero and started moving each time the basic pulse occurs. But one basic pulse must not follow the previous one until a sufficient interval has elapsed to allow the delayed or echo pulse to arrive. If this safety precaution is not followed, the delayed or echo pulse might not arrive until after the "second or third time around" of the watch hand; one could not be sure which. Thus the indicated time interval would be ambiguous to the extent of some integral number of whole time sweeps, or pulse-recurrence-frequency intervals. The situation is quite analogous to timing an interval with a stop watch that has a sweep second hand but no hand for recording whole minutes.

To avoid such possible ambiguities, the pulse-recurrence-frequency and, accordingly, the time sweep speed, must be decreased to an amount depending on the longest delays or distances expected to occur in the radio distance-measuring system. But slowing the time sweep decreases the accuracy of time measurement as previously explained.

6.4 PHASE-DISTANCE SYSTEMS (TYPES B AND C)

In absolute-distance or differential-distance systems using phase comparison of continuous waves, the same general connection between accuracy and ambiguity holds true. At an operating frequency of 100 kilocycles per second a wavelength is about 10,000 feet; a phase difference of as much as 10 degrees, for example, corresponds to only 280 feet. At 1000 megacycles, a wavelength is about 12 inches; 10 degrees of phase difference correspond to a mere ⅓ inch. As either round trips or differential trips are actually concerned, the difference in a line of position corresponding to 10 degrees phase difference may be half that just quoted. Thus, by choosing a suitable radio frequency, the phase method of measuring distances (as in Decca, Raydist, and related methods) can, in principle, be made very accurate. In the Raydist system, a phase type of absolute or hyperbolic system using heterodyne methods for increased precision of measurement, an accuracy of 1 inch to the mile is claimed.

The trouble, however, is that phase characteristics repeat themselves in cycles of 360 degrees per wavelength. A phase difference of 10 degrees is indistinguishable from one of 10 plus 360 or 370 degrees; and so on. Thus, the phase-measuring device or distance indicator shows phase differences from zero to 360 degrees only without revealing the total number of whole cycles or wavelengths involved. It is like having a very accurate vernier or micrometer scale without a main or coarse scale; or like trying to measure a long distance with a tape that is graduated in sixteenths of an inch but has no whole inch or foot figures printed on it.

One way out of this manifold ambiguity problem is to use a recording or cumulative type of phase meter, the instantaneous indication of which at any time is actually the algebraic sum of all phase changes occurring since a particular setting was made at the start of the trip, or at some other known point. This method is especially feasible if the navigational system is of such a type that the indications are given by a direct-reading mechanical meter. The Decca, a phase type of differential-distance system, uses this idea.

This scheme, however, has the fundamental drawback, for any navigational aid, of dependence on unbroken continuity and unfailing accuracy of all past readings. There must be no interruptions caused by propagation disturbances, such as static or severe fading; by temporary malfunctioning of transmitting or receiving equipment, such as blown fuses; or by human errors such as pulling a wrong switch.

Furthermore, one could not risk tuning out a station or pair of stations temporarily to get a fix by tuning to another station for an intersecting line of position. The airplane would have to carry two or more receivers and indicators for reading two lines of position continuously and concurrently for fix purposes. Navigation on a long trip could not be by piecemeal guidance from a succession of different stations of limited range along the route, as this would also require interruptions. A given ground station, or pair of stations, would have to be sufficiently powerful to serve an airplane at all points during its flight.

Or, as described in connection with ambiguities in other systems, the difficulties might be resolved by dead reckoning or other alternative means of location having requisite precision. One might use direction-finder bearings on the system stations or on any radio stations within range, for a fix sufficiently accurate for resolving the ambiguities. Direction finding may perhaps always be useful as an adjunct to other types of radio navigational aids, for rough checks, ambiguity resolution, emergencies, or other auxiliary purposes; but no navigational method which absolutely requires direction finding or other schemes to resolve ambiguities can be called an independent or a self-sufficient system. A navigational aid should be capable of giving a fix at any time "from scratch" no matter how lost the navigator may be, how careless he may have been at a past time, or what temporary interruptions may have occurred in the past.

II. Air Navigational Applications

7. *General Navigational Requirements*

7.1 Types of Information

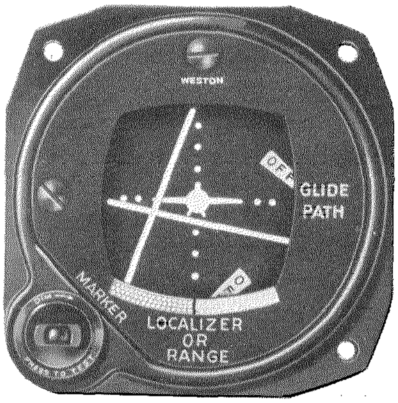
Navigation means to conduct a ship or airplane to its destination via a desired route. For the fullest and most convenient solution of the problems occurring in navigation, the navigator should have available:

A. Knowledge of his absolute position, in terms of latitude and longitude, bearing and distance, or any other intersection of lines of position. Aural, visual, celestial, and radio means are at hand.

B. Knowledge of altitude above the earth. As flight is three dimensional, this item is necessary for a really complete fix. Barometric and radio means are available. A surface ship would like to know the depth of the water below it, for which purpose sonic devices, the acoustic analogues of radio altimeters, are available.

C. Knowledge of his position relative to a prescribed flight path or intended track. One can use any line of position, whether established by visual, celestial, or radio observations, as a flight path, but directly indicated radio flight paths are most convenient.

D. Knowledge of the true heading of the airplane. A compass is necessary for this, radio means being at present insufficient, except possibly in a round-about manner, using a certain disposition of direction-finding beacons. A direction-finding transmitter at the north pole is one theoretical solution.



Photograph courtesy of Weston Electrical Instrument Corporation.

Modern type of cross-pointer meter or deviation indicator. Vertical and horizontal pointers, respectively, give left-right and up-down directions to the pilot for following a localizer (or radio range) and glide path. The small flags marked "Off" are hidden from view in normal operation; their appearance is a warning of lack of signals or other malfunctioning of the system. The marker-beacon lamp is located at the lower left corner of the instrument.

E. Knowledge of time. A fixed time reference is provided by a watch which may be corrected periodically by radio time signals which are related to Greenwich time.

F. Knowledge of speed. True ground speed may be determined if (*A*) and (*E*) are always available, but no single instantaneously reading instrument is available for this purpose. Certain approaches, however, are possible, as by using rate meters on a radio distance indicator, or by using the Doppler effect.

G. Knowledge of the presence and location of other moving (i.e. non-charted) objects nearby. Radar is the indicated radio aid for this anticollision service. Knowledge of weather conditions ahead might be construed as falling within this requirement. Weather conditions may be reported verbally by radio from ground stations; or anticollision type radar might even be used to give direct warning of certain types of weather conditions immediately ahead.

It is seen that radio, or electrical devices in general, are capable of providing the chief items of navigational information, at least in principle. At present, however, not all such devices have been developed to the requisite degree of accuracy, reliability, range, convenience, or practicability.

7.2 Reliability

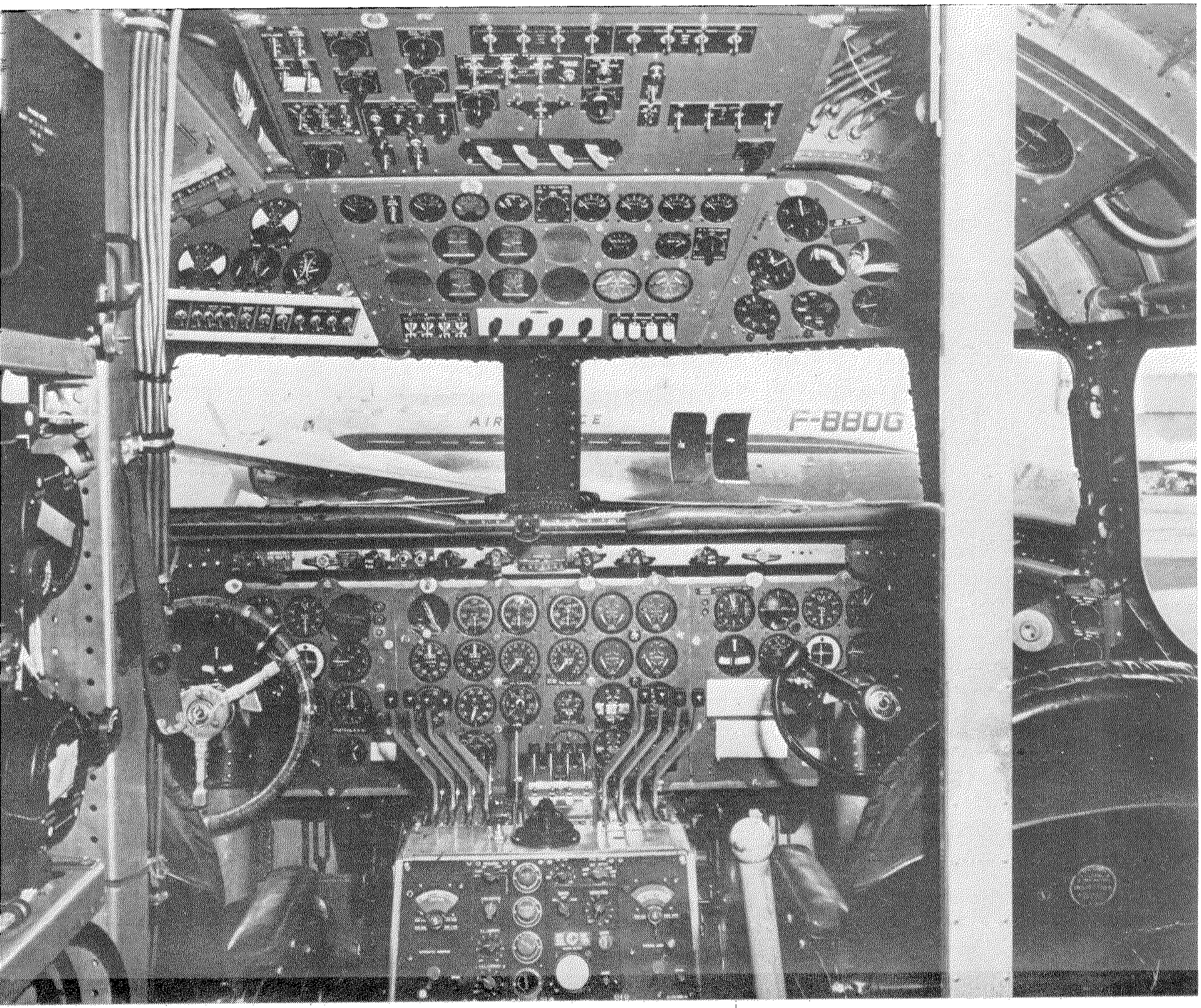
Reliability, in a radio navigational system, has two aspects. First, since transmission over distances is concerned, there is the matter of propagational reliability. The signals *must come in* at all times and over all specified service areas with sufficient strength to override interference and noise. Propagational reliability is a problem around which the general features of a system must be designed at the outset. It involves matters such as distance coverage, frequency, bandwidth, polarization, static, transmitter power, receiver sensitivity, interference, and ionospheric effects, and is an especially important problem in the long-range navigational case.

The second aspect of reliability is the conventional one that enters whenever life and limb have to depend on electrical or mechanical devices. In the case of airplanes, even more so than in the case of sea or land carriers, which can stop and wait things out if they are totally lost, extremely high standards of reliability are required. Meeting these standards is especially difficult because of the size, weight, power, and drag limitations for airborne equipment.

In addition to the insistence on reliability for any airplane equipments, another safety feature has come to be universally demanded for airborne navigational devices. That is the provision for automatic indication of failure. It is recognized that no device can be 100-percent reliable forever; tubes may blow or wires may break in the face of the most rigid manufacturing and preflight tests. As a result of such a failure in a navigational device, a navigator may be totally lost. But bad as that may be, one thing that is definitely worse is if the navigator does not know it and trusts the equipment, flying on blindly in his happy ignorance. If, at least, he is apprised of the fact that the device is out of order, he may be able to repair the defective part, radio for help, or use other means of navigation.

In this connection, there are two points of interest. The first is that those navigational systems that give visible indications on a cathode-ray tube have a certain advantage. They may be more bulky, more tiring to read for long times, and require time-consuming manipulations before a reading is made; but on the other hand they allow visual monitoring of the actual *quality* of the incoming signals. This tells a great deal about the propagation aspect of reliability and something about the electromechanical functioning, but it is not a cure-all.

Pilot's compartment of a Douglas DC-6 airliner. An important group of instruments supplies navigational information and includes altimeters, compasses, and radio range and landing indicators. Standardization of appearance and placement of the newer radio navigational indicators for safest and most effective use is an important subject being studied by various coordinating committees.



The second point is that emergencies will always occur and preparations must be made for them. Thus, it is highly advisable for airplanes to have alternative navigational aids, even if less accurate or convenient. A sextant can always be carried, of course; a direction-finding receiver or attachment, as an alternative radio aid, appears to be highly advantageous also. In some systems—as in most directional transmitting (type-D) systems—a direction-finding attachment could be provided to work off the same receiver and the same ground stations, with no extra ground station equipment and insignificant extra airplane equipment. In any case, in emergencies a direction finder can be used with a communications receiver to take bearings on any stations within range, such as broadcasting, radiotelegraph, or surface-ship installations.

Reliability is not to be confused with accuracy. The closeness with which a navigator can pinpoint his location on the earth is one thing. It may be within ten feet or within ten miles depending on the system. High reliability means that whatever the limits of error are, they are known in advance and will not be exceeded. What limits of error might be desirable, or sufficient, depend on circumstances.

7.3 Accuracy

One might first demand 100-percent accuracy for a radio navigational aid; but this might be obtained only at the expense of a number of other factors of importance. Fortunately, something less than 100-percent accuracy may be all that is absolutely necessary, so that in practice some sort of compromise between the various factors is sought.

A fundamental law is that the total amount of information that can be transmitted by a radio system per unit time depends on the total amount of radio channel space or bandwidth used. This is a law of far-reaching implications in communications work, instances of it being familiar to all radio engineers. It is of equal fundamental importance in radio navigational aids, although in this case the amount of information required per unit of time is much less.

Total information per unit of time is a product compounded of such terms as amount of information given out in each direction, area of service, accuracy of information, freedom from ambiguities, quickness of indication, and convenience of presentation. Engineers can improve one or another of these factors by good design, but there comes a limit, depending on the bandwidth, beyond which an improvement in one factor is at the expense of deterioration of another. And the supply of available bandwidth is limited since it must be shared by many users without mutual interference.

The physical relation between two of these factors, accuracy and ambiguity, has already been described. Some proposals suggest getting around this complication by providing, in essence, two systems side by side: one a coarse system, good enough for quick work and for resolving ambiguities; the other a very accurate or fine system. Similarly, some proposals for improving the coverage of an essentially non-omnidirectional system entail construction of two installations, suitably oriented to fill in each others' gaps.

Such solutions involve increase in the total bandwidth used, of course, but may be valid solutions since the accuracy-ambiguity relation is one of the most troublesome among the factors mentioned. But for a practical single system, using a bandwidth dictated by other considerations (propagational, legal, etc.), one must decide which of the two properties is more important, razor-sharp accuracy or complete lack of ambiguities; or else make a reasonable compromise. For certain special military applications, which like atomic bombs one hopes may never be used, pin-point accuracy may be essential. Everyday civil air navigation is something else.

The type of mental gymnastics required to unravel complex ambiguities might intrigue a professor of logic, a cross-word-puzzle expert, or other "armchair navigator." The crew of a big airliner completely lost at sea, with time and fuel fast running out, might favor a navigational device that tells them directly and positively that they are *somewhere* within a circle 10 miles around Bermuda, rather than a device that tells them they are *exactly* 12.863 miles due east of Bermuda, but possibly also 104.394 miles due south of Halifax or 0.875 mile due west of the Empire State Building.

The distance that an airplane may be off course without serious danger or inconvenience, the time allowable for making readings and playing around with ambiguities, and the likelihood that the airplane may at all cross zones of ambiguity, depend on the exact purpose of the radio navigational aid.

In landing systems, for example, the need for precision is greatest; being "off" by 100 feet, even less, might well be fatal. Also time is short so that indications must be very quick and direct. A complete fix dare not be given by methods requiring successive tuning to different stations and reading different instruments; all the necessary information must be given concurrently. On the other hand, omnidirectional or wide coverage is not necessary; is not even desired. The pilot is supposed to keep on one unidirectional beam, once he gets started on it; thus he runs little risk of being so far off as to "slip a cog" or end up on a wrong line of position even if ambiguous lines were around somewhere.

To get him safely started on the correct beam while far out, other navigational aids that are less accurate but give wider coverage and less ambiguity are depended on, after which the airplane keeps strictly on the one landing beam. In short, in landing systems the highest precision is necessary; but is feasible because the airplane's route is such as to avoid situations where ambiguities are significant.

In cross-country medium-distance nagivation, on the other hand, provision should be made for omnidirectional lines of position and fixes. As the pilot cannot fly a whole trip by sticking to one beam, and may have to tune to different stations for a fix, ambiguities do have to be considered. By the time the pilot has tuned back to the original station, after getting an intersecting line of position from a second station, he may have "slipped a cog" or moved across one or more zones of ambiguity, such as lobes, wavelengths, pulse-recurrence intervals, etc., without knowing it. Therefore, ambiguous zones should at least be a certain distance apart, depending on the speed of flight and on the reading time of the navigational aid. In short, in overland ranges, ambiguities should be totally absent or at least fairly wide apart, which is feasible as the need for precision is not extreme. The navigational facilities need only enough accuracy to guide airplanes into airport areas and to landing beams far out from the airport where they are wide, and to segregate traffic safely into lanes perhaps 500 or 1000 feet apart. For this application, there are possibilities of using radar-type systems which give the requisite accuracy and yet are omnidirectional, free of ambiguities, and sufficiently quick-reading.

In very-long-distance transoceanic service, traffic is less dense and there is more room and time available, so that ambiguities or slower reading times could be tolerated to increase accuracy. In this connection, it is to be noted that the total bandwidth limitation for the long-distance case is more stringent than for the short or medium cases because of propagational requirements; very great accuracy could not be obtained without excessive deterioration of other desirable factors. Great accuracy, fortunately, is least necessary in the long-distance navigation case. All that is necessary is the ability to follow a definite flight path at midsea with a precision of, say, 10 to 20 miles; and the ability to make a fix within about the same distance to initiate a "square search" for disabled craft in emergencies. In directional-type systems, this precision amounts to an angular precision of approximately one-half to one degree, at a maximum range of 1500 miles. For the same angular precision, the lateral distance accuracy increases proportionately as the ground station is approached; thus the beam, even though it may be fairly wide far out, brings the airplane using it for track guidance close in to the ground station, like a funnel. In any long-range navigational aid, relatively poor accuracy is permissible at midsea, where most of the time exact knowledge of position is merely of academic interest; closer in, however, the precision should at least permit a reasonably close landfall or permit the airplane to strike somewhere within the service area of a more accurate medium-range radio navigational facility, extending say 100 miles offshore, which can guide it to a specific airport or radio landing beam.

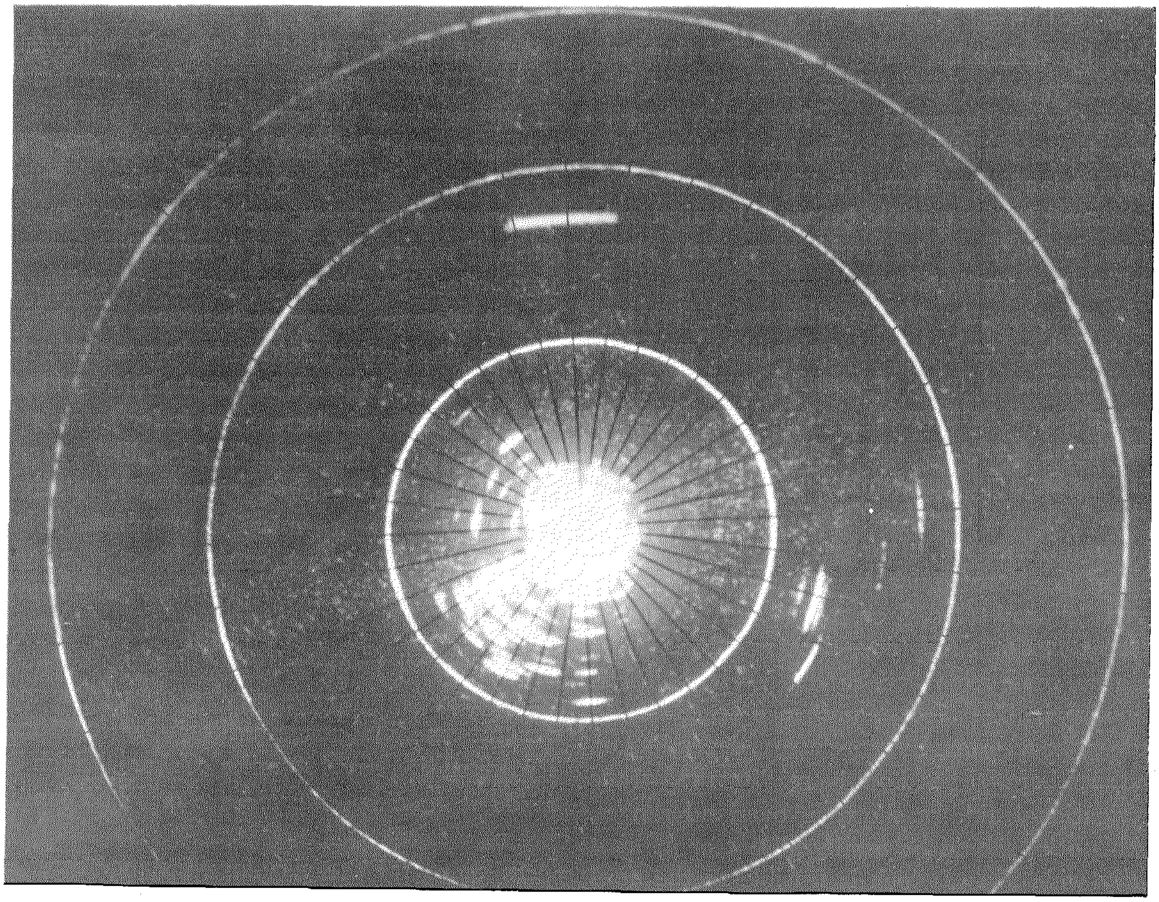
In short, for the very-long-distance case, the total bandwidth limitation is severest so that all factors must suffer; but fair amounts of ambiguity and inaccuracy can be tolerated in practice.

7.4 Presentation

The way in which the navigational indications are presented is not of basic importance, compared to reliability and accuracy, but it is something that interests the navigator; or would especially interest the second pilot, radio operator, or flight engineer, in the event that navigators are dispensed with even on long-distance flights. It is more important on aircraft than on surface ships because airplanes do not have the time and space facilities for long drawn-out observations and calculations; that indeed is one of the reasons why radio navigational aids are at all favored over celestial navigation, even when visibility conditions are good.

Automatic direct indications, requiring nothing other than pushing a button to receive a desired ground station, are of course ideal. Meter or dial indications are direct; radar plan-position displays are direct and even pictorial. At present, cathode-ray screens are not quite bright enough for viewing in a daylight-illuminated cockpit; but this may be remedied in time. Cathode-ray screens are more tiring on the eyes of observers than are meters, especially when protracted observations are necessary. Cathode-ray tube indicators, especially of the radar plan-position type, are at present rather bulky pieces of equipment for a cockpit. It is conceivable that graphic automatic position plotters, simulating a plan-position display but more suitable for daylight viewing, could be provided, if desirable, with meter-type indications. For this, two lines of position would have to be indicated concurrently.

Certain types of cathode-ray indicators, such as used in pulse-timing systems, require considerable time before a reading can be made, but



View of plan-position-indicator screen showing combined normal-radar and assisted-radar displays. Range circles are at 5-mile intervals. Wide pip at top is beacon response from an airplane. Prominent pips at lower right are natural echoes from structures on the ground. Other illumination is due to ground clutter, permanent echoes, and random noise. With the natural-radar receiver off, the screen becomes completely dark except for the bright airplane beacon response.

have the advantage, already mentioned, that the quality of the signals is visually apparent at all times. Meter indicators are quickly read, but must always be provided with an automatic warning feature to show that the meter and everything behind it is in working order. Meters also have the advantage that the mechanical motion of the pointer can more feasibly be transmitted to operate an automatic pilot, if desired.

Aural indications have the great advantage of not requiring special indicating equipment, the airplane's regular radiotelephone receiver being used and, incidentally, providing continuous monitoring of the quality of the signals. This may be an advantage for pilots of small private airplanes; but aural indicating systems appear to be on their way out. The indications are tiring, and are either quite approximate or time consuming.

The necessity for quick, convenient, and direct indications depends on the particular purpose of the system, being, as already described, most pronounced in landing systems and least urgent, though still desirable, in long-distance navigational aids. It should be remembered, moreover, that other things being equal, the simpler the type of presentation and the simpler the skills required in making various receiver and indicator adjustments, the less likely are ordinary human errors and careless mistakes.

7.5 INDEPENDENCE

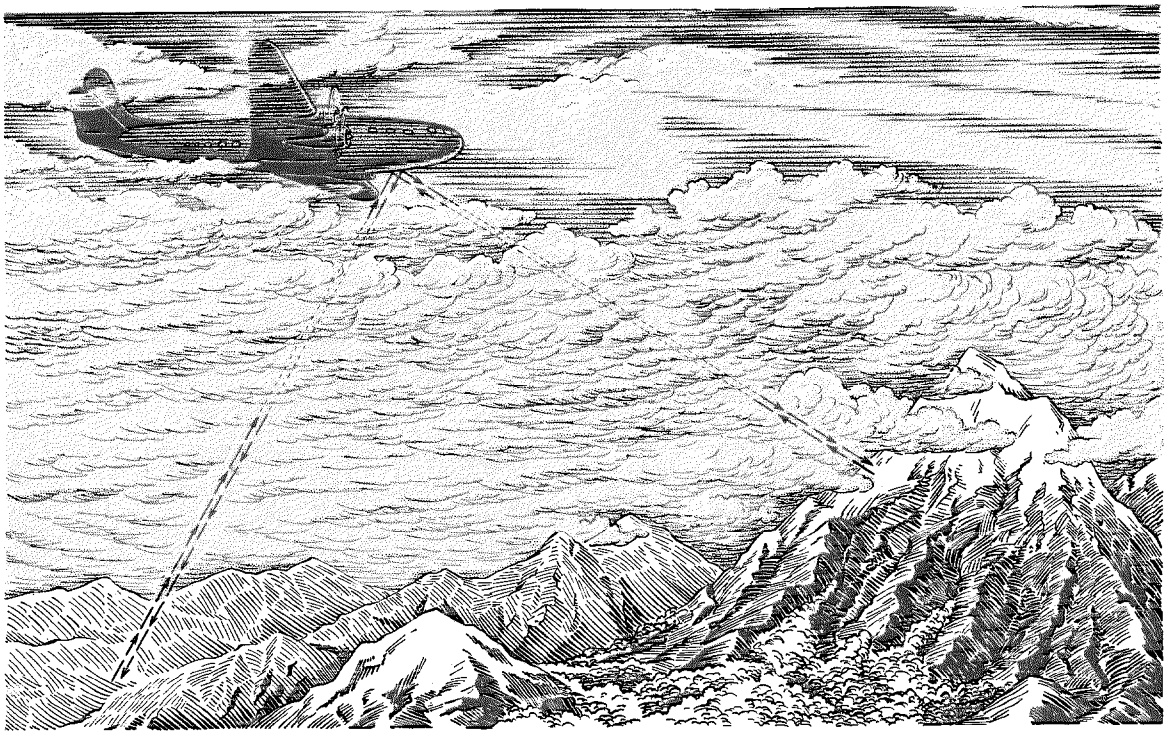
Self-sufficiency or independence has been stressed as a fundamental requirement for radio navigational aids, regardless of type of service. This means that there is no need for reading any additional instrument, like a compass, to determine a line of position or a fix; and that the indications are given directly on the airplane, instead of occurring on the ground for a second party to relay to the airplane.

Table I shows how each of the four basic types of radio navigational aids compare on this requirement. In each type, two possible designs have been separately considered: a design in which the basic or primary transmitter is

TABLE I

CHECK-LIST OF BASIC RADIO NAVIGATIONAL SYSTEMS

Basic Type of System	Bearing Systems				Distance Systems			
	Directional Receiving		Directional Transmitting		Absolute Distance		Differential Distance	
Primary Transmitter Located on	Ground	Plane	Ground	Plane	Ground	Plane	Ground	Plane
Examples	Radio Compass, Aural and Automatic Visual Direction Finders	Ground Direction Finder, Huff-Duff	CAA Ranges, Consol, Sonne, Navaglobe, Rotating Beacon	Directional Part of Airborne Radar	Oboe	DME, Shoran, Distance Part of Airborne Radar	Gee, Loran, Decca, Popi	Micro-H
Independent of Plane's Compass	No	Yes	Yes	No	Yes	Yes	Yes	Yes
Original Observations on Plane	Yes	No	Yes	Yes	No	Yes	Yes	No
SYSTEM MAY BE SELF-SUFFICIENT	No	No	YES	No	No	YES	YES	No
Number of Radio Links Required for Establishing One Line of Position Including Links for Speech, Synchronization, Natural Echoes, and Active Response								
Ground Station to Plane	1	1	1	1	2	1	2	1
Ground Station to Ground Station	0	0	0	0	0	0	1	1
Plane to Ground Station	0	1	0	1	1	1	0	2
Total Number of Links Required	1	2	1	2	3	2	3	4
SUITABLE FOR LONG RANGE AS NO EMISSION REQUIRED FROM PLANE	YES	No	YES	No	No	No	YES	No
YES ON BOTH COUNTS			√				√	



Radio altimeter measures absolute height of the airplane by timing echo signals reflected from the terrain below.

airborne, and another in which that transmitter is on the ground.

From the first half of Table I, it is seen that systems of only the following types are fundamentally capable of being independent: ground-based directional transmitting, ground-based differential distance, and airplane-based absolute distance.

It appears, incidentally, that an airborne radar installation cannot be a completely independent navigational aid. The absolute-distance part is self-sufficient, but the directional-transmitting part gives relative direction, requiring a compass. A ground radar, considered as a navigational aid, is also incomplete. The directional part may or may not be, but the distance indications have to be relayed back to the airplane.

Even among the three preferred types of systems, different specific types of radio aids appear necessary for short-, medium-, and long-range navigation because of the different characteristics of various types of radio emissions and because of the different kinds of practical problems concerned. For the short-range applications such as landing systems, the chief problem is the extreme precision of guidance required. The medium-distance applications for overland navigation and traffic control present a great problem in the multiplicity of indications that must be provided and the large number of airplanes that must be cared for simultaneously. In the long-range case, traffic density is negligible and great accuracy is not a primary requirement, the chief problem being that of providing reliable signals at the large distances involved.

A brief outline of some of these special problems and a discussion of some present and future solutions follows, all based on the principles outlined under the four basic types of radio navigational aids.

8. Short-Range Applications

8.1 RADIO ALTIMETERS (TCI)

Radio methods that are used for measuring absolute distance (type-B systems, Section 3) are applicable to making a determination of the vertical height or altitude of an airplane. As they indicate the height above the terrain immediately below the airplane, not above sea level, and without regard to barometric weather fluctuations, radio altimeters are also called *absolute* altimeters or terrain-clearance indicators (TCI).

The ground immediately below the airplane reflects radio energy emitted from the airplane. Any of the forms of type B may be used: pulse timing, continuous-wave phase comparison, or saw-tooth frequency modulation. Two special problems have to be taken into account. First, short distances have to be measured with a sensitivity at least equal to a barometric altimeter, 10 feet or so. Any one of the three radio methods described results in ambiguities which limit the maximum altitude range of the system. Often, two range settings are provided, a low-altitude high-sensitivity range using high pulse-recurrence frequencies and fast time sweeps, and a high-altitude lower-sensitivity range using a low pulse-recurrence rate and slower time sweeps.

Secondly, one must be sure that the echoes used for measurement are those returned from ground points directly underneath. Fixed beacon responders obviously may not be used in radio altimetry, nor are they necessary for adequate signal strength for the distances are short and the natural reflecting surface is large. To ensure that the echoes come from straight below, the transmitting antenna must have a certain amount of downward directivity. Although no scanning or motion of the antenna is needed, this directivity requirement does increase the size and drag over that of a plain antenna. The indicator can be a cathode-ray tube or meter.

If the transmitter were fixed on the ground, an airborne responder might be used; the distance indicated would not be the vertical terrain clearance, but the slant range. Also, the information would have to be relayed back to the airplane. Any conventional ground radar, in fact, measures the slant range of the airplane. The vertical height may be computed from this by trigonometrical formulas or devices if the angle of elevation of the airplane is known. This is done, in some military applications such as height finders, by using a sharply directive antenna that can scan up and down and noting the vertical angle at which the echoes appear.

It should be noted that regardless of the state of perfection of radio altimeters, they may supplement conventional barometric altimeters but not supplant them. A radio altimeter, or terrain-clearance indicator, gives indications with respect to the terrain below and is thus useful for avoiding collisions with large fixed objects such as mountain tops. Its indications are not suitable for establishing a system of assigned altitude layers for safe segregation of traffic. An airplane in perfectly level flight over rough country would have constantly changing terrain-clearance indications, which would differ from even those of nearby airplanes flying at the same actual level. For traffic regulation and avoidance of collisions between airplanes, altitude indications should be with respect to a datum level that is the same for all airplanes near each other. "Pressure altitude" indications, as given by a barometer-type instrument, meet this requirement because the reference datum is sea-level pressure; this is actually variable with weather conditions but the variations are common to all airplanes in the same general region.

In this connection, there is a quite recent air-navigation principle known as "pressure-pattern" flying, which makes use of information gathered simultaneously from radio and barometric altimeters. Comparison of absolute and pressure altitudes enables the wind factor to be determined, which in turn increases the accuracy of dead-reckoning procedures. An extension of this method theoretically allows airplanes to fly the shortest *air* course between two distant places, the point being that, because of wind effects, the direct great-circle route is not necessarily the route of minimum flying time and fuel consumption.

8.2 Landing-System Requirements

Instrument approach or "blind landing" systems require basically one fixed straight-line track; thus ground-based directional transmitting systems (type D, Section 5) are in order. The antenna array, on or near the runway, must be small; the range need not be long, 10 to 20 miles or so; but the line of position must be exceptionally precise. For all these reasons, landing systems use very high radio frequencies.

What have been heretofore referred to as *lines* of position are actually parts of *surfaces* of position. Ordinary antennas radiate energy

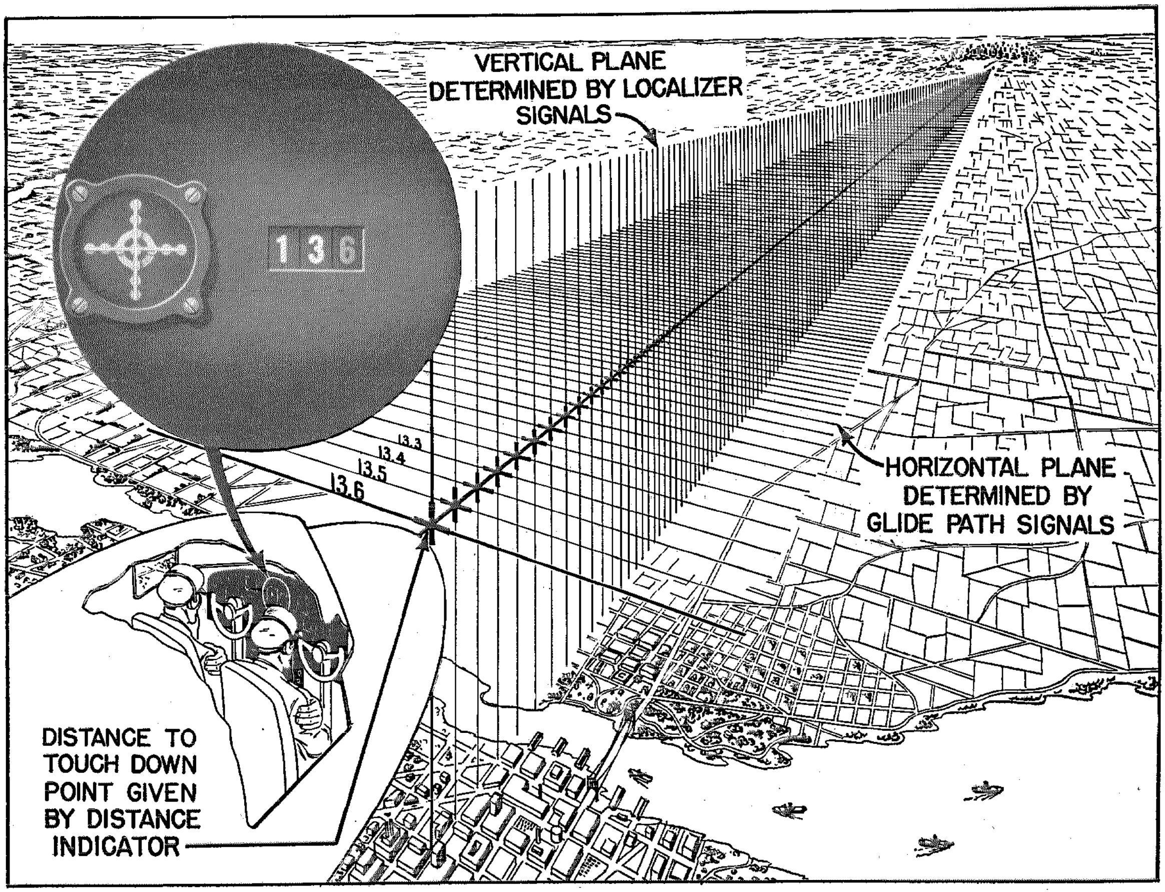

VERTICAL PLANE DETERMINED BY LOCALIZER SIGNALS
1 3 6
13.3
13.4
13.5
13.6
HORIZONTAL PLANE DETERMINED BY GLIDE PATH SIGNALS
DISTANCE TO TOUCH DOWN POINT GIVEN BY DISTANCE INDICATOR

Fig. 10—Instrument landing system principle used in Navaglide.

in all directions, vertically as well as horizontally. Thus, for example, in an absolute-distance system a given echo-time measurement means that the observer is somewhere on a spherical *surface*, of corresponding radius, centered on the ground station like a giant dome. The intersection of this spherical surface with the earth's surface gives the conventional *line* of position. For most navigational purposes, however, one is concerned with only that *line* on the earth's surface; this is because the altitude of an airplane is negligibly small in comparison with its horizontal distance from the ground radio stations. In directional transmitting applications considered up to now (except in the angle-of-elevation application) the surfaces of position are vertical plane surfaces, whose intersections with the earth's surface produce the conventional straight radial lines or great circles of position.

In landing systems, a true line and not a surface of position must be furnished, for now the altitude of the airplane and the proper gradual change of it are of vital concern. This true line of position should be a straight one rising at a certain small gliding angle from the desired touch-down point on the runway, and should be aligned with the runway.

8.3 Instrument Landing Systems (ILS)

Examples of instrument landing systems are the glide-path localizer such as the so-called Civil Aeronautics Administration Indianapolis and Army SCS-51 systems,[9,10] types in the development of which Federal played an active part; and other equipments developed by Sperry, Bendix, Massachusetts Institute of Technology, and other organizations. Two intersecting sur-

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[9] H. H. Buttner and A. G. Kandoian, "Development of Aircraft Instrument Landing Systems," *Electrical Communication*, v. 22, n. 3, pp. 179–192; 1945.

[10] S. Pickles, "Army Air Forces Portable Instrument Landing System," *Electrical Communication*, v. 22, n. 4, pp. 262–294; 1944.

faces of position are created by two directional transmitting arrays, horns, or dishes, properly oriented. Generally the switched-lobe equisignal principle, as described under type-D radio ranges (Section 5), is used for extreme precision of each surface of position. Each lobe transmission has a characteristic tone modulation, thus enabling a receiver to separate the two and compare their strengths.

As seen in Fig. 10, the localizer transmissions result in a vertical plane surface of position aligned with the runway. The glide-path transmissions result in a plane surface of position slightly inclined up from the horizontal. The intersection of the two plane surfaces creates now a *true line* of position, which constitutes the landing flight path or beam. On the airplane, both vertical and horizontal deviations of that airplane from this line of position are indicated automatically and visually by the cross-pointer or left-right and up-down meter. There are two pointers, one for the equisignal glide path, the other for the equisignal localizer path. If the pilot is off the beam laterally or vertically, the corresponding signals are out of balance and the pointer for that direction moves off center. In doing so, it indicates the direction in which to steer to get back on the beam.

For safest and smoothest landings, especially if completely automatic landings are contemplated, a complete three-dimensional fix should be given at all times and must include position along the beam. In the Navaglide system being developed by Federal out of its earlier landing systems, a direct-reading indicator giving the distance to the touchdown point is included on the airplane. This is of the pulsed absolute-distance (type-B, Section 3) principle, the responder-beacon being situated at or near the touchdown point.

It is conceivable that distance along the landing beam could be indicated by a precise directional transmitting station (type D, Section 5) off to one side of the runway; but in view of landing speeds of airplanes, the indications would have to be direct, quick, and simultaneous with the beam indications. Conceivably, also, the information could be provided by sensitive barometric or radio altimeters. An airborne direction-finding indicator (type A, Section 2.2) might also be used in conjunction with one or more transmitters located to one side of the localizer course, and in fact this system of compass locators has been used as a supplement to or substitute for marker beacons along landing beams.

8.4 GROUND-CONTROLLED APPROACH (GCA)

In this so-called talk-down system, both the distance of the airplane and its deviations from the prescribed landing path are determined by essentially ground-radar means of high precision. This type of landing system was developed by the Radiation Laboratories at Massachusetts Institute of Technology. Equipments were manufactured for wartime service by Gilfillan, Federal,[11] and Bendix. The system is now being adopted for civil use. Extended surfaces of position are not created, but only small angular sectors of the surfaces as shown in Fig. 10. Two antennas, very sharply directive in directions at right angles to each other, are used. These are relatively small radar antennas or parabolic dishes, and are scanned through a small search or exploring angle, one left and right, the other up and down, from the landing path. One of these two type-D (Section 5) directional transmitters (both using pulses) is also used as a type-B (Section 3) absolute-distance system. If the airplane is on or not too far off the beam, the directional antennas, during their scanning or searching, will detect it and also measure its distance. The position of the airplane relative to the prescribed path is displayed pictorially on a cathode-ray screen. These observations, however, are made on the ground. The information concerning the airplane's deviations from the beam, its distance, and corrective steering instructions, must be relayed to the pilot over another radio link, usually by radiotelephony. Such procedure may be advantageous, especially in emergencies, as it is possible to guide to a landing an airplane which has no special equipment other than a radiotelephone receiver. Also, the ground-controlled-approach system allows the ground observers to see any other airplane that is near the landing path at the time.

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[11] J. S. Engel, "Landing Aircraft with Ground Radar," *Electrical Communication*, v. 24, pp. 72–81; March, 1947.

Since the sectors of search about the beam are small, and the airplane has no detecting mechanism, it is very unlikely that the airplane could, coming in from a distance, get started into the beam by itself. Accordingly, a conventional omnidirectional search radar of the plan-position-indicator type is used to spot the airplane far out and guide the pilot into the beam; after which the more precise radar takes over. The search radar information must also be relayed to the pilot.

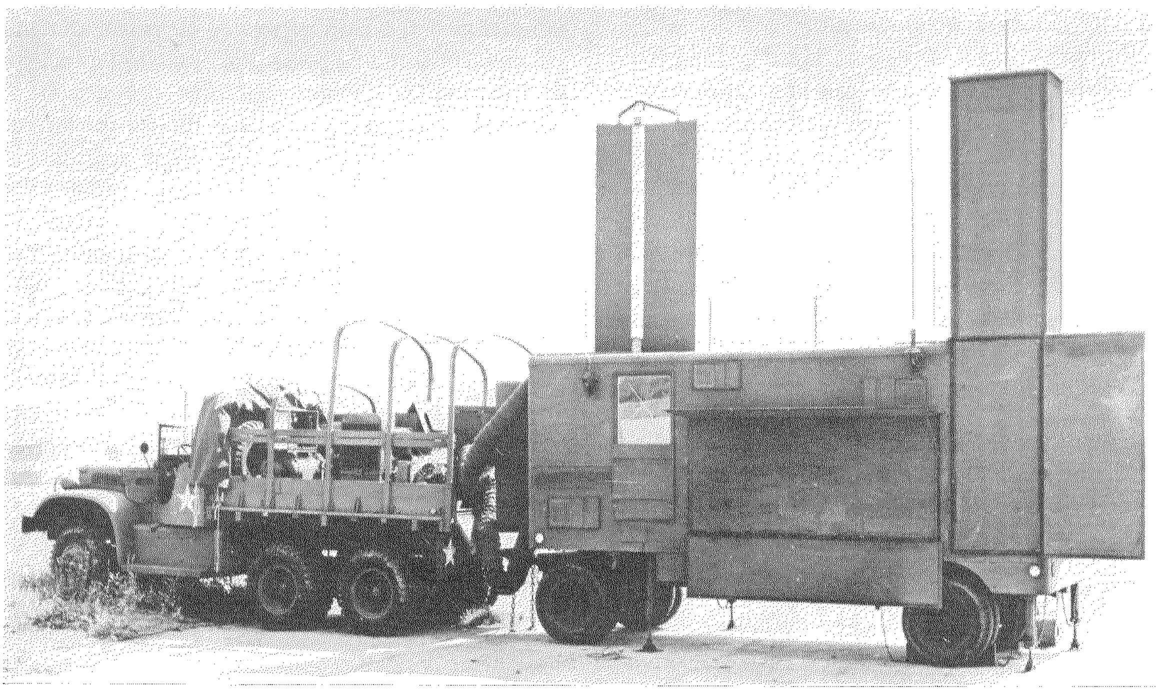
In instrument landing systems, the surfaces of position are quite broad and the airplane has direct indicators. Thus, even if he should enter the airport area quite far away from the landing beam, the pilot may be able to locate the beam by himself. Or, knowing in advance the bearing of the beam, he could guide himself into it by using other medium-range navigational facilities. At least, the pilot will known immediately and directly when he comes somewhere near the beam by observing his own cross-pointer.

The relaying of information for a ground-controlled approach might possibly be done by other nonverbal methods, but the system still requires an additional radio link and intermediary personnel to make the observations on the ground, just as in ground-based direction finding. Thus the ground-controlled-approach system is not an independent or self-sufficient navigational aid.

8.5 Radio Mileposts (Marker Beacons)

Because of unavoidable errors in observational methods and in electrical and mechanical devices, no navigational indications are 100-percent accurate. Knowledge of position is thus always inexact to some extent, even when systems and devices are well engineered. A surface of position thus is actually a thin shell or wedge; a line of position on the earth is actually a narrow strip of position; a fix is not a mathematical point but a small area or solid *volume of position*, its size corresponding to the amount of uncertainty.

Aside from all this, it is possible intentionally to devise a directional-transmitter installation so that it produces a fairly extended and definite volume of position. This principle is used in so-called marker beacons (fan, Z, cone of silence). The aural or visual indication is of the simple on-or-off type, and the pilot knows only that the



Wartime model of a mobile ground-controlled-approach (GCA) installation. Power and air-conditioning equipment is in truck at left; radar and communication antennas are seen on the roof and side of the trailer unit. The trailer contains all equipment and indicators required to enable the operators to "talk down" an airplane.

airplane is inside or outside of a specified region.

The marked space may be shaped like an inverted cone with its vertex resting on the ground station, or like a thick fan standing on end as in the fan type. It is effectively limited horizontally by the directional properties of the antenna array (type D, Section 5), but being limited in height by transmitter power and receiver sensitivity, it is indefinitely high and intercepts airplanes at any likely altitude.

These radio markers may be coded for identification to inform the pilot of the zones through which he is passing, the system being similar to that of mileposts or buoys for surface traffic. In air navigation, the markers have been used to provide rough indications of important points along radio ranges such as the entrance to airport control areas, intersections with other ranges, or for local warning purposes. They are thus a sort of radio curtain, used for much the same purpose as the curtain of wires (telltale) before the entrances to railroad tunnels or low bridges. A series of two or three marker beacons is also used to give knowledge of position along landing beams of the instrument landing type. Marker beacons for all these purposes are essentially makeshift devices, are on their way out, and will no doubt disappear completely when omnidirectional radio ranges are installed for cross-country use so that a radio fix is possible at any point, and when landing systems include automatic distance indicators.

Marker transmitters may perhaps be retained for ground observation and control of air traffic. Automatic responders in the airplane might be tripped when passing through the marker region and report to the ground coded information revealing identity, altitude, direction, and time of the transit. This would be analogous to the block signal in railroad traffic organization, where the trains trip switches along the tracks and cause indicating lights to flash in the dispatcher's office.

9. *Medium-Range Applications*

9.1 General Requirements

For overland traffic service, ranges of about 100 miles or so are sufficient since chains of stations may be installed to cover larger areas or distances. Convenient direct beams between important air terminals are desirable, as is omnidirectional coverage, so that navigators can use radio aids for both track guidance and fixes at any location. At present, the most extensively used radio aids in the U.S.A. are the Civil Aeronautics Administration's aural A-N four-course ranges, but these will be supplanted by improved types which are visual, omnidirectional, or both. There is some use of airborne direction finding, and this will probably continue; even ground direction finding is used somewhat. For the ranges required, medium radio frequencies may be used, but the trend seems to be toward higher frequencies where antennas may be smaller, power less, static less troublesome, and accuracy greater.

This trend is heightened by the exhaustion of available channel space in the medium- and high-frequency bands. The still-higher frequencies offer more room for expansion, and the line-of-sight propagation limit permits frequency assignments to be repeated at fairly close geographic intervals without danger of interference. Hence the current intensive developments in the 960–1215-megacycle band, allocated in the U.S.A. for many future air-navigational and traffic-control aids, largely of the pulse type (wide band). Further raising of the frequency, except for specialized aids, appears questionable, for at appreciably higher frequencies there arise considerable problems in respect to tuning, frequency stability, transmitter power, and atmospheric absorption.

Ambiguities are undesirable because the coverage should be omnidirectional, flights may be short, and traffic quite dense. While there is no need for the extreme precision that is required in landing beam systems, high frequencies may be used for aircraft navigation for the altitude at which airplanes fly extends the line-of-sight considerably beyond the sea-level-horizon distance. It is extended to about 150 miles for an airplane at 5000 feet. Thus radar or allied systems, with their very small antennas, pulse transmissions, and other advantages, may be put to good use.

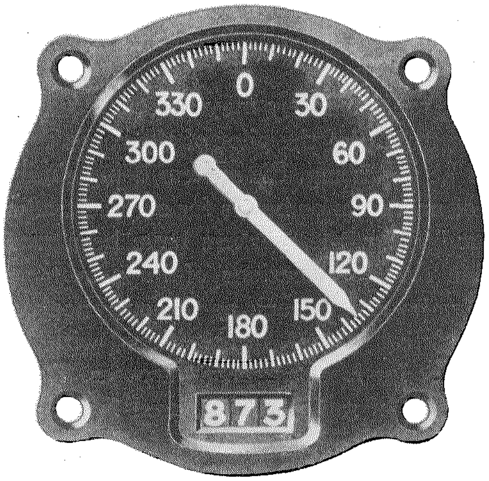
9.2 Distance Indicators (DME)

Distance indicators are absolute-distance (type-B, Section 3) devices, similar in principle with radio altimeters, and may likewise use pulse, continuous-wave, or frequency-modulated transmission. The range is greater than for radio altimeters, extending out to possibly 100 miles, or as far as the power or line-of-sight propagation allows. The accuracy need not be so great and, consequently, ambiguities are not prevalent. The indicated distance is again the slant range; but the horizontal distance is generally so great, compared to the height of the airplane above the earth, that the two are practically equal.

Unless the airborne distance indicator is part of a complete radar plan-position indicator, a ground responder-beacon is necessary. Distance indicators which are not part of a radar are generally of the meter type, and therefore require stronger pulses than do cathode-ray indicators. They give no directional information. The location of the reflecting landmark may be definitely established by using responder-beacons which either code the aided echoes, or respond on a definite radio-frequency channel, or both. The use of a special transmission frequency for the responder-beacon permits echoes from other reflecting bodies to be disregarded by the airplane receiving apparatus as in assisted radar. Responder-beacons are obviously saturable if a number of airplanes make simultaneous use of their facilities. This number can be increased by giving each airplane's interrogation signal a distinctive pulse-recurrence rate. The receiver must then include not only a time-measuring device but also an automatic "tracking" mechanism (strobe) for recognizing the desired echoes from those put out by the ground beacon in response to challenges from other airplanes.

Distance measuring equipments (DME) are very desirable operationally, as distance to destination may be read directly without the need for plotting a fix by crossed lines of position and then scaling off the distance on a map. The service is limited to moderate distances by the requirement for airborne transmitters; the practical limitation is actually to radio line-of-sight distances because the short pulses required for accurate time measurements necessitate the use of very high radio frequencies.

The various national and international policy-making bodies in aeronautics such as the Civil Aeronautics Administration, Provisional International Civil Aviation Organization, and Army Air Forces have recognized the urgent need for automatic meter-type distance indicators to make possible a complete distance and bearing (Rθ) navigational system in conjunction with existing or proposed directional-type aids. Development of distance-measuring equipment operating in the 960–1215-megacycle band is in intensive current progress by Federal, Hazeltine, and other organizations. Equipments were demonstrated at the October, 1946, session of the Provisional International Civil Aviation Organization at Indianapolis, and are scheduled for extensive flight testing preparatory to commercial and military installation. The associated directional system which, with the distance-measuring equipment, will provide a complete polar-coordinate or Rθ navigational system along airways, cross country, and around airports, is to be the new Civil Aeronautics Administration omnidirectional range system operating in the 112–118-megacycle band.



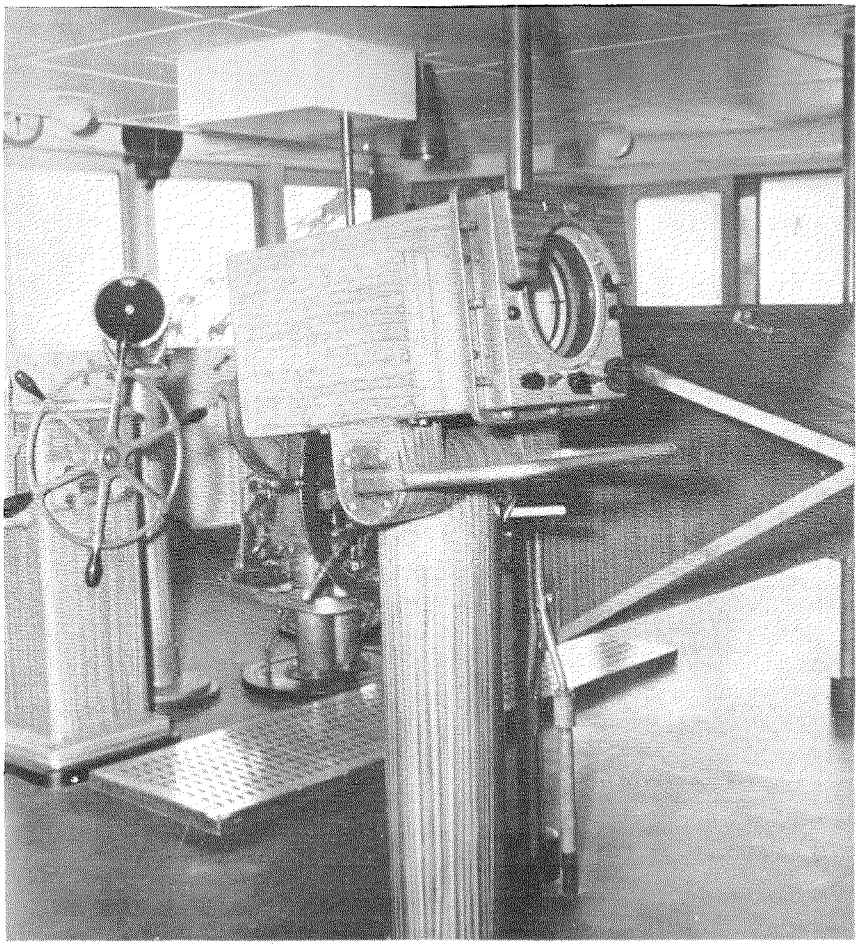
Simple meter type of presentation for an Rθ radio navigational system. Distance in miles and tenths is shown by the number-wheel type of indicator; bearing or azimuth is shown in degrees by the pointer. As distance and bearing are indicated simultaneously from the same radio landmark, complete information for a position fix is constantly presented in direct terms.

9.3 Radar for Navigation and Anticollision

Radar essentially gives a pictorial display in the form of spots of light on a screen and reveals the location of distant objects with respect to the position of the radar station. This information could be used in reverse to ascertain the location of the radar station itself, if the true positions of the distant objects appearing on the radar display were known.

The first use is well fitted for military purposes, spotting objects which the enemy may prefer to conceal. With responder-beacons on airplanes to increase the range or clarity of service, the use of radar for civil aviation as part of the problem of ground observation and control of air traffic is very promising. The return pulses from the airplane could even be coded to reveal, always or on demand, the altitude, identity, or similar information regarding the airplane. Use would be made of the principle of identification-of-friend-or-foe (IFF) systems of Section 3.4. For this purpose, radar is a direct and independent system; the basic transmitters are on the ground, indications are in convenient pictorial form, and various elaborations and refinements are possible. The use of radar for a number of observational and automatic reporting functions may ease the load on pilots and on radiotelephone channels for routine reports, and possibly, by remote indicating systems, the control tower may avoid use of radiotelephone channels for routine orders to airplanes.

The second or reverse use is as a navigational aid. A ground station obviously needs no such aid; but a ship or airplane might use radar to locate itself by reference to known landmarks, such as coded responder-beacons, harbor buoys, coastlines, bridges, or other natural features appearing on the display. Navigation then might be very skillful, even in close quarters. Several important considerations, however, enter the picture. If the radar is ground based, observational information must be relayed back to the airplane; therefore the system is not an independent one. If the radar is airborne, a compass may still be necessary because all airborne radar bearings are relative to the axis of the craft. (If *two* or more known landmarks could be depended on always to appear on the plan-position indicator, true directions could be determined.) Finally, a radar installation requires heavy bulky rotating directional antennas; thus mobile radar, though quite practical for surface ships, appears for the present to be impractical for most civil aircraft.



Photograph courtesy of Raytheon Manufacturing Company.

Mariners Pathfinder shipborne radar installation, an all-weather aid in marine navigation. Marine radar is being increasingly used. Sets are generally of the short-range type, showing nearby objects with great accuracy, so that other ships, buoys, and obstructions may be located.

This is unfortunate because radar as a navigational aid can do one thing that no other system can do. It can show the presence and location of other moving objects for anticollision warning service. Knowledge of one's own position alone, no matter how exact, is inadequate for this purpose; the need for anticollision service is becoming more and more urgent with increasing density of air traffic. The fact that the airborne plan-position display shows the position of other objects relative to the axis of the airplane is just what is wanted. Limited airborne radar for anticollision use only and not providing a complete navigational service is less impracticable. Low transmitter power will do as the range need only be several miles or so, very sharply directional antennas are not required, and only a small sector ahead of the airplane must be scanned; thus space, weight and drag requirements are less burdensome.

The physical movement of an antenna to cover all desired directions is an advantageous feature of radar and allied systems. By making the antenna very sharply directional, the transmitted power is concentrated instead of being spread out, thus increasing the effective radiated power and range of service. The rotational rate of the antenna, however, must be slow enough to permit a sufficient number of pulses to hit each reflecting object. The pulse rate might be increased to insure this, but at the expense of increased transmitter power.

9.4 ROTATING-BEACON PRINCIPLE (ORFORDNESS RANGE)

The uniformly rotating unidirectional ray feature which is characteristic of plan-position-indicator radar can be applied in other ways, even at comparatively low frequencies. In one design already described, the Civil Aeronautics Administration omnidirectional range (Section 5.4), which is a medium-distance method and may possibly be applicable also for long distances, the uniform rotation is produced electrically rather than by physical means. Before these methods were invented, the use of actual physically rotating antennas of large size, operating at medium frequencies, was tried.

An example of this type of directional transmitting system is the British installation at Orfordness. A fairly large directional antenna radiates a rather broad beam. The whole antenna is bodily rotated at a fairly slow rate, say once per minute. Each time the radiation is to the north, a distinctive reference or marker signal is emitted. With an ordinary stop watch, the observer measures the time elapsed between reception of the north signal and of the rotating beam. Knowing the rotational rate, this time interval is a measure of the angle through which the moving beam has turned from the north until it momentarily points at the observer, and gives the observer's true bearing from the ground station. The stop watch could be calibrated directly in degrees of azimuth.

A fully automatic indicator could also be devised. Other interesting forms may suggest themselves. The rotating directional transmitter might have a modulation changing uniformly with time so that its instantaneous value corresponds to the instantaneous bearing. Different audio-frequency tones might be used for this purpose. The direction of the antenna might even be recited by phone modulation every five degrees or so, each observer hearing only his bearing. Systems of this sort were used during the war as rough homing aids for airplanes. This method is very simple in principle but, at medium frequencies, antenna systems are either insufficiently directional to be accurate or much too bulky for physical rotation. Thus this system has not been extensively used.

9.5 NAVAR PRINCIPLE

That same principle, however, is promising for use at higher frequencies, where antennas may be sharper in directional characteristics, smaller, and capable of faster rotation so as to give quicker indications. Furthermore, the principle opens up an interesting attack on the problem of integrating the ground traffic-control problem with the navigational problem, for medium-distance overland traffic.

One would start with a conventional ground radar station of the plan-position-indicator type for ground observation of air traffic. This type of radar always has a physically rotating, sharply unidirectional antenna as its basic element. To extend this into an omnidirectional and unambiguous rotating-beacon type of radio range, practically all that is needed is the additional emission of a distinctive omnidirectional north signal, each time the radar beam passes through north.

This is the principle of the omnidirectional range or azimuth-meter service which is one part of the Navar system being built by Federal. The timing is done automatically and true bearings or azimuths are indicated visually and directly on the airplane by the pointer of a 360-degree-type meter.

The other basic navigational part of the Navar system is a direct-reading distance indicator on the airplane. This uses pulses sent out by the airplane, as described under distance indicators (Section 9.2). The responder-beacon is at the ground station; its transmitter, in fact, may be the same one that produces periodic omnidirectional north signals in the direction-indicator system. A large number of airplanes may use this simultaneously without interference by use of the random-pulse-recurrence principle, also described in Section 9.2.

The Navar system resembles radar but is not identical with it; Navar and radar may work together. In radar, both the distance and directional transmissions and indications are on the ground. In Navar, the directional transmissions occur at the ground but their indications are produced in the airplane; the distance transmissions, which require no bulky directional or rotating antennas, take place from the airplane and the indications occur directly in the airplane. Moreover, a large part of the equipment, both on the ground and in the airplane, may be used in common for the bearing, distance, ground plan-position indicator, and possibly other functions.

Like radar, Navar gives a complete fix by means of a bearing line of position and a distance line of position, both on the same landmark and both indicated directly and concurrently. Furthermore, both indicators may be small easily read meters instead of bulky and not-too-bright cathode-ray tubes. In fact, for the medium-distance overland navigational problem, the presentation is practically the same as the ideal instrument shown in Fig. 1, except that indications are in terms of bearing and distance rather than latitude and longitude. This is even preferable for land orientation, where latitude and longitude are not particularly meaningful at first glance. Where air terminals are the important reference points, bearing and distance from them are most pertinent. The Provisional International Civil Aviation Organization has recommended that overland air navigation be standardized according to a distance and bearing system of reference. This implies a combination of some form of airborne distance-measuring equipment and omnidirectional azimuthal service.

For track guidance, any straight-line radial path leading to or from the ground station could be flown with facility by keeping on a line of position of fixed bearing; any circular flight path, such as an orbiting or holding maneuver, which is common near airports in preparation for landing procedures, could be easily flown by keeping on a line of position of fixed distance. A left-right meter or an automatic pilot could be used for guidance of the airplane along such paths, or even along any desired offset straight-line path, if a computing device which worked off both bearing and distance indications simultaneously were provided. Development of such computers is in progress by Minneapolis-Honeywell, Collins, and other organizations. Fig. 11, shows useful types of flight paths possible with a distance-bearing system.

9.6 Automatic Position Plotting (APP)

It is conceivable that the Navar type bearing and distance indicators could be provided with a mechanical polar-coordinate tracer to indicate the position of the airplane graphically and continuously on a map centered about the region of the ground station.

Any radio navigational system that indicates two intersecting lines of position concurrently gives a continuous indication of fix. The two lines of position might be radial as in any type-D system (Section 5), or as determined by combining direction finding (type A, Section 2) and compass indications, or they might be two circular lines of position from an absolute-distance (type-B, Section 3) system, or even two hyperbolic lines from a differential-distance (type-C, Section 4) system. In all such cases, it is conceivable that the two simultaneous indications could operate an automatic position plotter. This has been proposed by Bendix for the direction-finder and compass case.

The bearing-distance combination, however, has a number of advantages over any other combination of lines of position. First, the bearing-distance indications are from a single landmark and may occur from use of one radio-frequency channel, thus requiring only one tuning operation. Bearing and distance from a single place, even as numerical quantities, give a clearer mental picture of a location than do two bearings, two circles, or two hyperbolas, from two landmarks. Finally, the airplane needs to be within range of only one station, rather than two widely separated, specially related stations at the same time.

The automatic position plotter would be mechanically much simpler with the polar coordinates of bearing and distance than with two bearings, circles, or hyperbolas, especially as in the latter cases the distance and direction between the two landmarks would not be any fixed amount. A manual adjustment would have to be provided for introducing these factors into the apparatus, depending on the relative location of the available pair of ground stations. Also, the two ground stations might not always appear on one sectional map. Finally, with crossed straight lines, circles, or hyperbolas, fix errors are liable to be greatly magnified under certain circumstances, as when the two lines of position are not close to being perpendicular; whereas distance and bearing lines of position from one landmark are always at right angles, the best angle of intersection for high accuracy.

9.7 Relayed Radar (Teleran, Navascope)

Radar, together with the advantage of giving simultaneous bearing and distance lines of position from one landmark, can give pictorial plan-position displays and is uniquely suited for anticollision service. Airborne radar, however, has the installation drawbacks already noted. The idea has therefore been proposed to use a ground radar station to collect such information as distance and bearing of airplanes and terrain features for display in conventional form for ground observation and control of air traffic. All this information might then be relayed to the airplane by radio for navigational purposes. The airplane would require a relatively small radio installation without bulky directional and rotating antennas and would require no powerful transmitter.

This scheme would seem to allow one powerful ground radar station to serve a large group of airplanes with navigational and anticollision information cheaply. The information preferably should not be relayed to the airplanes verbally; the airplanes should be provided with either a picture or a duplicate of the original ground plan-position display.

The picture method has been proposed by the Radio Corporation of America as the essence of its Teleran system for air-traffic control and

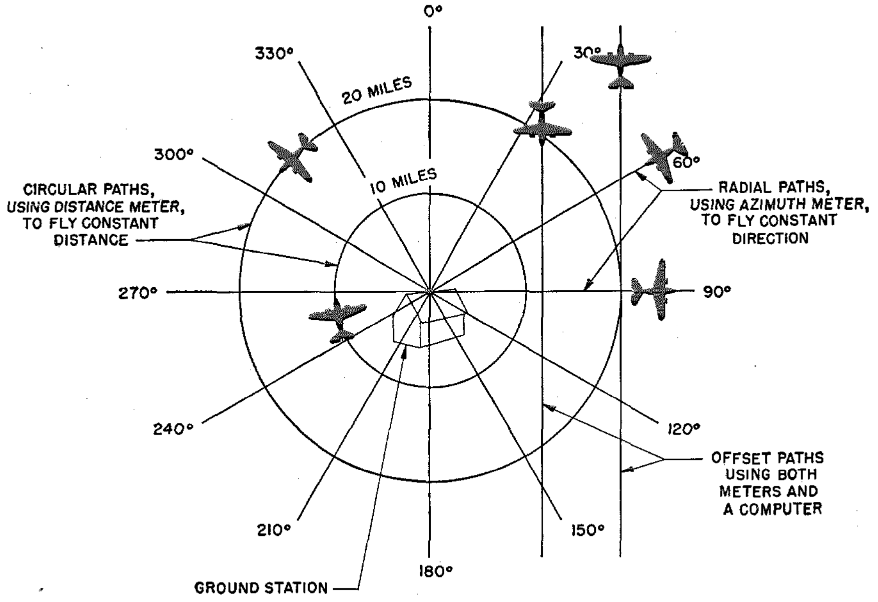
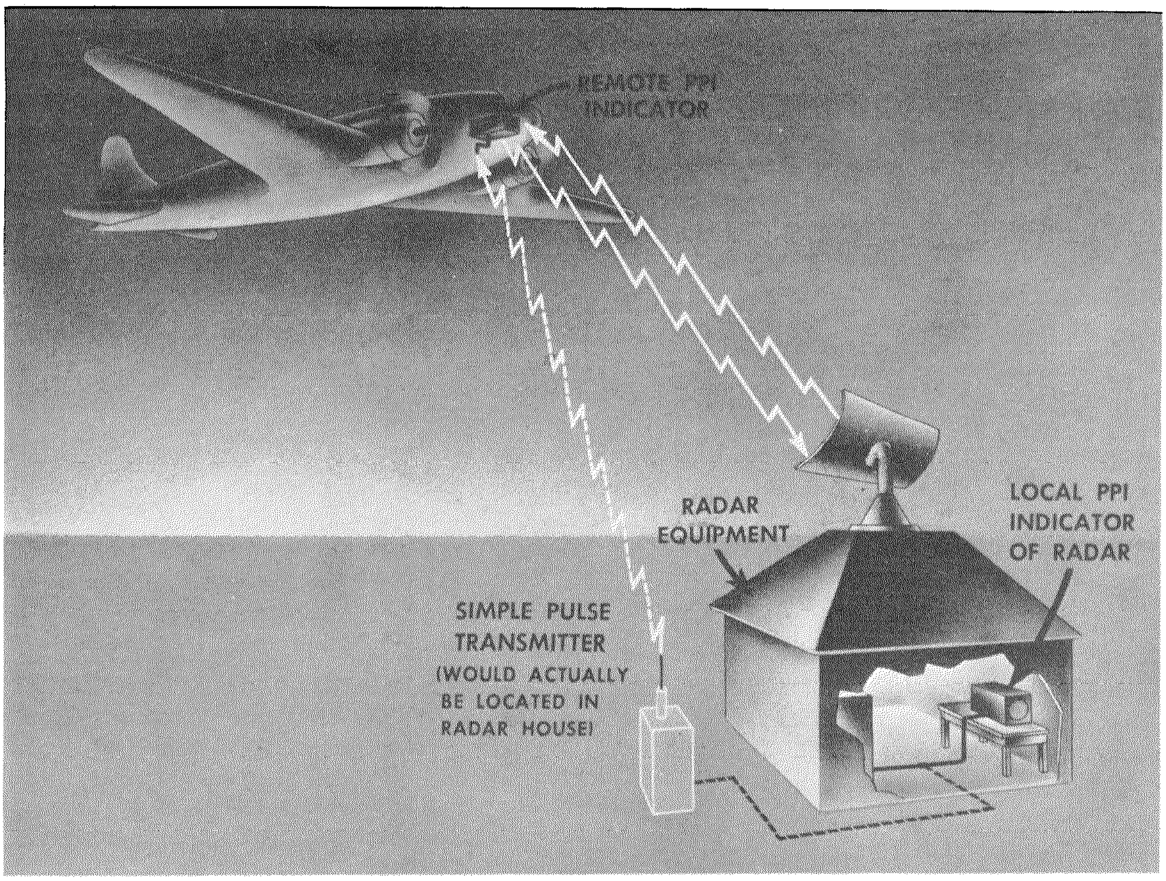

0°
30°
60°
90°
120°
150°
180°
210°
240°
270°
300°
330°
20 MILES
10 MILES
CIRCULAR PATHS, USING DISTANCE METER, TO FLY CONSTANT DISTANCE
RADIAL PATHS, USING AZIMUTH METER, TO FLY CONSTANT DIRECTION
OFFSET PATHS USING BOTH METERS AND A COMPUTER
GROUND STATION

Fig. 11—Useful types of flight paths possible in an Rθ or distance-bearing radio navigational system. Tracks may be accurately followed by reference to a distance-azimuth meter like that shown in a previous illustration.



Principle of remote radar as used for the Navascope, or airborne duplicate of the ground radar plan-position display.

navigation. Principles of ordinary television practice are to be used. The ground radar display is produced in the usual manner; the plan-position indicator is scanned by a television camera and broadcast by an omnidirectional television transmitter. The picture is displayed on an airborne television receiving set. This scheme offers a number of interesting possibilities; for example, additional information, map indications, traffic orders, etc., could be written on the plan-position indicator or held in front of the television camera, and would instantly appear on the airplane television receivers.

The duplicate or remote radar method has been proposed by Federal as one part of its Navar system for traffic control and air navigation. Again a conventional ground radar station produces a plan-position display in the usual manner. Airplanes are given a duplicate of this display in much the same manner that an additional display would be produced in an adjoining room, except that radio instead of wire links are used. The incoming echo pulses are used not only to create the light spots on the ground plan-position indicator, but are broadcast in original form but on another radio channel in all directions. Other control pulses for synchronizing the time sweep and the bearing sweep are likewise broadcast, all in proper time relationship, so that the airplanes use essentially the same pulses to create a plan-position display as does the ground station and in the same manner. Certain extra information could also be put in at the ground station by special electronic means. Also, the pilot can vary the content and manner of presentation, such as scale, heading, or offset, to suit his purposes.

This type of relayed radar, which has been used for certain wartime applications involving the repeating of radar displays between surface ships and airplanes, is proposed as part of the integrated Navar system to give a pictorial display on the airplane in addition to the primary meter-type indications of distance and bearing. In this connection the remote radar or Navascope display would use, to a certain extent, the same ground and airborne equipment installed for the airborne distance and bearing indicators as well as for the ground radar display.

While relayed radar, whether televised or remoted, appears to obviate the need for a complete airborne radar, on closer analysis three fundamental defects appear.

In the first place, the system is a "second hand" one as indications originally assembled on the ground are relayed to the ultimate user over an extra radio link, whereas independence has been stressed as a desirable feature for any navigational aid.

Secondly, the original radar display uses assisted techniques; cooperating equipment, such as responders, are required on all airplanes, and only those so equipped will appear on the display. However, until airborne radar is made more practical, relayed radar may be a useful compromise solution to the anticollision problem.

Thirdly, the information given by relayed radar alone is necessarily incomplete; the system is not a self-sufficient navigational aid.

The ground station collects only directional and distance information regarding the airplanes. Airplanes are not identified; they are just so many spots of light on a cathode-ray screen, indistinguishable from each other. When the ground display is transmitted to the airplane, the center of the picture is the ground station and the spots of light still look alike. The result is that the observer on a particular airplane, looking at the radar picture, does not know which of the many spots of light corresponds to his airplane. The only way he could identify his own light spot is by knowing his own bearing and distance, the information he is trying to obtain. Attempting to resolve this vicious circle is like trying to lift oneself by one's own bootstraps.

The fact that the ground radar antenna acts as a rotating beacon, irradiating an airplane once during each rotation, could be used to determine bearing. This principle is used to actuate the direct-reading azimuth indicator which is one of the basic parts of the Navar system. Other airplanes, however, might be at the same bearing at the same time.

Additional schemes may be used to aid in judging which spot of light corresponds to a given airplane. If, however, an independent source of distance information were also continuously available, then the question would be answered completely. If desired, the bearing and distance information could be combined automatically to mark in some distinctive manner the light spot corresponding to the observer's airplane. This principle is incorporated in the remote radar part of the Navar system, for independent airborne distance and bearing indicators are already provided as primary navigational devices.

For complete identification and really effective anticollision warning service, the relayed radar display should give altitude information also. The observer, for anticollision purposes, does not need to know the actual altitude of his or other airplanes; he need only know which other airplanes are near his own airplane. He does not even need to see on the display those airplanes which are at considerably different altitudes. By the following principles, in either televised or remoted radar, this might be accomplished at least effectively enough for the purpose.

The ground radar display originally shows all airplanes regardless of altitude; but each airborne relayed display is devised so as to be altitude selective. To accomplish this, the aided echoes from the responder-beacon in the airplane are coded in accordance with a barometric altimeter mechanism. The coding thus corresponds to the instantaneous altitude of the airplane. This feature permits each airplane to include in its radar or television display only those spots of light indicating airplanes whose return pulses have the particular altitude coding corresponding to any altitude layer selected by the pilot.

Furthermore, in remoted radar there is the possibility of an automatically adjusted altitude presentation, that is, the barometric coder can control the relayed radar indications so as to show only airplanes at the same altitude region as the observer's airplane and this display level will follow the airplane as it changes in altitude.

Only *true mobile* radar, however, may be a first-hand, pictorial navigational aid and anticollision indicator. For anticollision service, it is unique and is the ultimate device. For practical reasons, mobile radar is only suitable at present for surface ships and possibly for very large aircraft. If anticollision service or a pictorial radar display is nevertheless wanted on airplanes, it might well turn out that some sort of relayed radar is the practical solution, especially for small or moderate-sized airplanes where weight, power, drag, and expense must be held down. For fundamental reasons, however, an independent self-sufficient navigational system, one with unambiguous observations made directly on the airplane, is still desirable as a primary navigational aid, with or without relayed radar as an adjunct.

10. Long-Range Applications

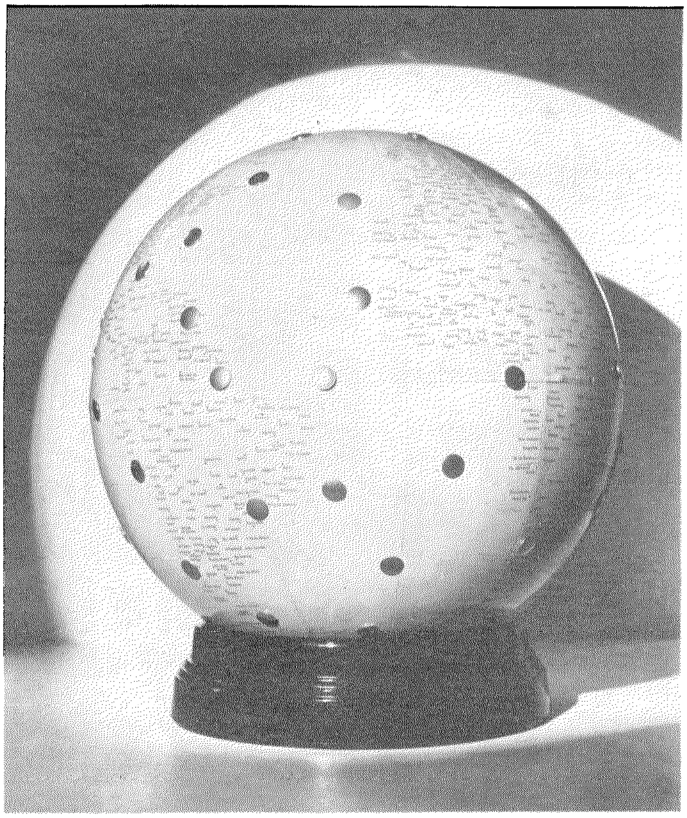
10.1 General Requirements

Navigational aids for transoceanic routes or routes passing over large desolate areas like the polar regions present special problems. Relayed radar, for example, is out of the question, for this facility requires chains of ground stations of limited line-of-sight range. If anticollision service should be desired, pure airborne radar is a necessity. As regards the main navigational items, great accuracy is not too important. Reliability of service over great distances is the big problem; in midocean there are no emergency landing fields. Midocean landings, even for seaplanes, are not too safe; and the trend seems to be toward land-type planes for transoceanic traffic. Good track guidance is desirable since on a long trip much time and fuel may be wasted by erratic wanderings off the direct course.

10.2 Distance Requirements

The first question that occurs when discussing radio navigational systems that are to be ultimately capable of covering all oceans, polar areas, etc., is just what distance range is required for the ground stations. It may be assumed that to establish a fix by crossed lines of position, an airplane must be within range of at least two ground stations whose lines of position cross at a usable angle; and that ground stations are to be at actual land points, coastal or island, and not on moored ships or man-made islands. A recent study[12] indicates that if a reliable range of about 1500 miles is postulated, sufficient land sites are available. With this range, 4 to 6 stations could service the important North Atlantic area, and some 60 stations could cover all oceanic and polar areas. For differential-distance type systems (Section 4), where for each family of lines of position two widely separated suitably related sites must be found, the situation is more complex, especially at midsea where islands may be few.

At 1500 miles, any radio navigational aid that requires transmissions of any sort, including echoes or responses, from an airplane appears to



Base map copyright by Rand McNally & Co., Chicago, Ill., R.L. 4728.

Globe view showing possible locations of ground transmitters of 1500-mile range, for providing radio navigational service at all points in the North and South Atlantic oceans.

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[12] P. R. Adams and R. I. Colin, "Frequency, Power, and Modulation for Long-Range Radio Navigation System," *Electrical Communication*, v. 23, pp. 144–158; June, 1946.

be out of the question, because an airplane cannot carry an adequate transmitter and antenna installation. The various basic types of radio systems are analyzed with respect to this requirement in the lower half of Table I. It is seen that only airborne direction finding, ground directional transmission, and ground differential-distance types may meet this condition. But among these, only the last two also meet the requirement of self-sufficiency or independence as analyzed in the first half of Table I.

This does not mean that *all* directional transmitting (type-D, Section 5) and differential-distance (type-C, Section 4) systems necessarily meet all the requirements; this depends on specific system details. It does mean that in looking for a radio navigational system that meets the two basic requirements for long-range use, no other types of systems need be considered. The question of which of the two types, or which specific system of either type, is the best or the most promising, can only be answered after thorough technical study of such details as accuracy, reliability, frequency and bandwidth requirements, availability of suitable sites, ambiguities, shape of lines of position, convenience of indications, degree of omnidirectional service, weight and complexity of equipment, and so on; and ultimately, only by actual test. Some general comparisons of the two favored basic types are next given.

10.3 DIFFERENTIAL DISTANCE VERSUS DIRECTIONAL TRANSMITTING

In a sense, these two types of systems favored for long range are different only in degree. In both cases, sets of two or more antennas are used and the emissions from the individual antennas must be accurately related. In both, the lines of position are necessarily hyperbolas because they result from a differential effect between two or more antennas. The hyperbolas, however, degenerate into sensibly straight lines at distances from the array that are greater than about 5 to 10 times the spacing between individual antennas; this condition holds true for all practical purposes in type-D systems. The line of demarcation between the two types of systems appears in several other ways. One may say that in directional transmitting systems the antennas are from a fraction of a wavelength to several wavelengths, say a few feet up to a mile, apart; while in differential-distance systems the antennas may be hundreds of wavelengths, up to hundreds of miles, apart. The latter condition has it advantages; as in trigonometric surveying and plotting, a long control base line results in more accurate indications, other things being equal. Also, errors caused by nonuniformity in the terrain in the immediate vicinity of the antennas become less serious.

In differential-distance systems, the signals from individual antennas are received independently and then compared in phase or in time. In directional transmitting systems, the signals from the individual antennas merge, so to speak, and only the net effect is observable; this net signal is compared with other net-effect signals produced by the same or nearby antennas.

In differential-distance systems, the antennas are so far apart that a radio link and very careful procedures must be used for synchronizing the two distant transmitters. The signals from the two antennas travel to the distant observer by quite different paths over the earth. In directional transmitting systems, the antennas are so close that a short wire link may be used for synchronizing the signals; generally, the identical transmitter is used for all antennas. On the other hand, the antennas in differential-distance systems may be of the simplest, nondirectional type, whereas in the directional transmitting systems the layout, polarization, modulation, balancing, and current relations of the individual antennas of the array must be carefully attended to, so that the correct directional patterns are always produced and are correctly modulated. All the signals, however, travel to the observer over a single propagation path.

The site-location problem is more stringent in type-C (Section 4) systems because two sites, suitably related, must be found to produce one set of lines of position. In certain regions, this might require raising the distance requirement, thus increasing the construction and power cost. Two sites mean two bases, services of supply, and groups of operating personnel. As Fig. 7 and the check list (Table I) shows, three distinct radio signal links per airplane are required for one line-of-position observation; this means fundamentally greater chance of breakdown.

All types of systems are subject to certain basic sources of error, which limit the accuracy of even the best-designed and engineered system. For very-long-range systems, radio propagation occurs via sky waves, that is, by waves which are reflected down to the receiving station by the various ionospheric layers. This always affects radio communication and navigational services. In directional systems, inherent errors may arise because of certain changes in the polarization of the reflected waves, and because of distortions of the wave front caused by certain terrain features. In any distance systems, inherent errors may arise because of deviations of the path of the radio signals from the most direct path, caused by irregularities in the ionspheric layers; in differential-distance systems, moreover, there are two distinct paths to the receiver, and each path may be affected differently. In all systems, errors may arise because of interference between signals arriving via the ground path and via the sky path or paths from the same transmitter. They may vary in time of arrival, phase, polarization, etc. Systems using very short pulses and cathode-ray tubes have an advantage in that the ground-wave pulse, arriving first, may be visually distinguishable from the pulses arriving over longer paths.

The comparative seriousness of the various inherent errors in the two types of systems, and the limiting magnitude of them in each system, have not yet been determined well enough to judge which type of system is preferable in this count, or to predict the limits of accuracy one may expect in long-range radio navigation in general.

The above sources of error are not in the transmitting or receiving apparatus but in the physical medium of propagation between the two, a medium over which engineers have little or no control. Outside of such inherent errors, which are characteristic of the basic type of system, in each specific system there may be errors caused by ambiguities, misadjustment, or mechanical malfunctioning, which errors may be reduced by good design and engineering and by careful and intelligent operation.

10.4 PROPAGATION PROBLEMS

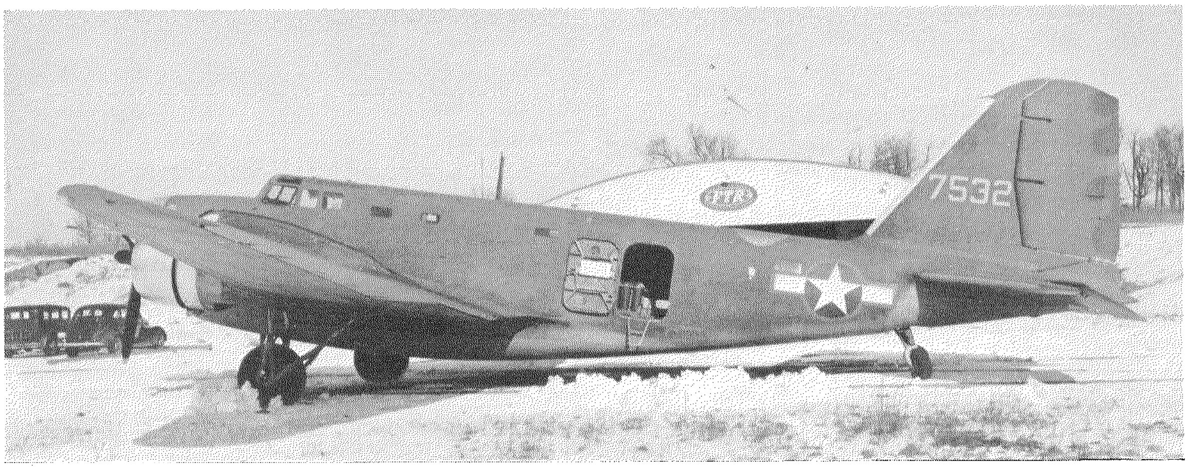
For short-distance applications, high frequencies are generally preferred because sharp directive effects are desired with small installations. For medium-range work, up to 100 miles or so, certain frequencies may be preferred for practical reasons, but almost any frequency would do. In both cases, the power requirements create no great problem. For a 1500-mile reliable range, however, the question of proper frequency for reliable transmission is a critical one as power, transmitter, and antenna costs are not inconsiderable.

The study[12] previously referred to has gone into these questions starting from first principles, and the results of the investigation prove to have some direct implications with regard to the suitability of specific systems for reliable long-range use.

For long-range propagation, only the low-frequency, very-low-frequency, and high-frequency bands may work. Medium frequencies, such as used for broadcasting, have very short ranges in the daytime and are variable at night. Very-, ultra-, and super-high frequencies are restricted to line-of-sight distances, about 250 miles for an airplane at an altitude of 20,000 feet.

High frequencies, as used in international broadcasting and communication, may travel long distances with comparatively little power and small antennas. Interruptions and disturbances, however, are very frequent and serious, the ionospheric layers being highly variable with respect to these waves. Fading is common and, at intervals that may be frequent and protracted, there are complete blackouts of reception, connected in some way with magnetic storms and sunspot activity.

Low and very low frequencies, as used since early days for very-long-distance communication, propagate quite steadily day and night, during all seasons and years, as the ionospheric layers that reflect them are less subject to upheavals. Also, one frequency serves for all distances, directions, and times of transmission, while for long-distance high-frequency communication, a transmitting station often has to choose among three or more available frequencies to fit the circumstances. Low frequencies are especially



Laboratory-equipped airplane used for gathering experimental data on performance of ground and airborne equipment during the development of a radio navigational aid (SCS-51 instrument landing system).

useful for radio communication in polar regions, where air traffic may be expected to increase in the future to take advantage of direct great-circle routes; in these regions high-frequency transmissions are subjected to such excessive ionospheric disturbances as to be practically useless, except for short distances.

Thus, from the all-important consideration of reliable and uniform propagation, low frequencies seem indicated for long-range navigational systems. Low-frequency transmitters, however, require very large antennas and considerable power. Also, the bandwidth should be extremely narrow for two reasons. First, static or noise, which may be very disturbing at low frequencies (less so in polar regions and more so near tropical areas), may be reduced in effect by use of narrow-band reception. Secondly, the low-frequency band does not have room for many or wide channels. Therefore, if a number of navigational stations are to be accommodated without interfering mutually or with other radio stations, their emissions must be so narrow in bandwidth as to use up very little channel space. The general conclusions of this study regarding the desirability of low-frequency and narrow-bandwidth operation for long-range radio navigational aids have found agreement in the recommendations and experiences of others. A great amount of experimental development work is being done on specific systems meeting these conditions.

In a subject with so many ramifications, accurate quantitative predictions cannot be made; but the aforementioned study concludes that for a 1500-mile service range, the greatest reliability may be expected with least power cost at a carrier frequency in the neighborhood of 80 to 100 kilocycles. The input power requirement, per ground station, for a receiver bandwidth of about 20 cycles, is stated as ranging from about 10 to 100 kilowatts, depending on the nearness of the station to the tropic zone. It is interesting to note that 100 kilowatts when translated into mechanical power is equal to about 140 horsepower or about one tenth the power output of a single modern airplane engine.

10.5 SPECIFIC SYSTEMS

Up to the present, no system satisfying all the basic requirements, let alone the features of convenience, of a long-range navigational system has been built. Some have been proposed or are being developed. In a number of cases, the system is merely an extension of a type used more or less successfully for medium distances. The difficulty is that mere increase of power is not always sufficient; the particular modulation, frequency, or other features being such that the systems are propagationally unsuitable for long-distance work.

The principal systems under discussion for long-distance use may, however, be mentioned and their long-range features itemized briefly. The first requirement is, of course, reliable service, which implies low-frequency narrow-band operation; then ambiguities, convenience, omnidirectional coverage, antenna complexity, etc., may be considered.

Loran and Decca are propounded as promising differential-distance systems, the hyperbolic lines of position and other general features being the same as already described under type-C (Section 4) systems. Standard Loran, which has had extensive wartime use, uses brief pulses, which require wide bandwidths and fairly high frequencies. Some promising work is being done on low-frequency forms of Loran; the inherent and practical problems (channel space, etc.) are considerable, and the final answer is not available. Decca, a recent development, uses pure continuous waves and is thus eminently suited for low frequencies and narrow bandwidths. The phase-comparison method, however, brings in the great problem of ambiguities which has been already described, and which may only be resolved by using recording meters, carrying double equipment for the receiving of two pairs of stations for a fix, and using extra high powers in the transmitters. Loran (at least in high frequencies) and Decca are capable of great accuracy, one requiring cathode-ray-tube observations and manipulations, the other requiring careful treatment of ambiguities. In Loran, although rather wide bandwidths are used, several stations can operate on the same radio-frequency channel without interference, through the principle of different pulse-repetition rates as described under distance indicators (Section 9.2). In Decca, even the two stations of one pair must operate on different channels so their signals may be distinguished at the airplane, where they are subsequently changed to identical frequencies for phase-comparison purposes. This requires essentially two receivers for observing one line of position and a special schedule of frequency allocations. Coverage is fairly wide so that many lines of position are available, though of hyperbolic shape and requiring special charts. In all cases, antennas may be the simplest nondirectional types, but accurate and reliable long-distance synchronization is required.

Among the directional transmitting systems propounded are Consol or Sonne, Civil Aeronautics Administration omnidirectional range, and Navaglobe. Consol and Navaglobe are inherently suitable for low-frequency narrow-band operation. The Civil Aeronautics Administration system, which was originally developed at high frequencies to replace the four-course A-N ranges, has special modulations that make the problem of adapting it to low-frequency narrow-band operation difficult, but some interesting work is being done on low-frequency designs of it. The Civil Aeronautics Administration and Navaglobe systems are omnidirectional and direct indicating, with little or no ambiguities. Sonne, which has been operationally used during the war, is capable of great accuracy, but the coverage is not completely omnidirectional, and alternative means, direction finding, for example, must be used to resolve ambiguities. Its aural indications require no special receiver but take considerable time. Sonne or Consol require an array of only two or three antennas; the omnidirectional Civil Aeronautics Administration and Navaglobe systems require five and three antennas, respectively. Navaglobe has not been in operational use, but an experimental model is under construction. In all type-D systems, the observations may not be as accurate as in differential-distance systems but should be precise enough for ordinary commercial purposes, and the antenna layout and excitation are critical. The lines of position are of the more convenient shape of great circles and require no special charts.

11. Conclusion

The foregoing descriptions of specific radio navigational systems were restricted to fundamental characteristics; in each system there may be hosts of other quite important considerations, intimate technical and other details, which further add to or detract from the suitability of the systems. Only tests over a period of years, especially for the long-range case, will give the final answer, insofar as any technical development can be considered final. In radio navigation, as in any other field, developments are not cheaply attained; Edison's well-known characterization of invention as being 10 percent inspiration and 90 percent perspiration still holds true. Credit for the ultimate developments in radio navigational aids, whatever form they may take, must be shared among the large number of individuals and organizations in many countries who have contributed to the advancement of the various branches of radio and electronics.

Frequency-Shift Keying in the I.T.&T. System

IMPORTANT international circuits of International Telephone and Telegraph Corporation associate companies are now employing frequency-shift keying for Morse and printer codes, and many others are scheduled for conversion to this method of operation during 1947. This method of operation makes possible utilization of printer and time-division-multiplex equipments on international circuits without increasing the power of the transmitters. On circuits where manual transcription methods are retained, frequency-shift provides the same circuit efficiency with an appreciable reduction in radiated power.

Historically, frequency-shift transmission was first employed with the old Federal-Poulsen arc transmitters. As they could not be keyed on and off, signaling was accomplished by changing frequency. This was referred to as back-wave or compensated-wave keying. However, no use was made of the back wave at the receiving point; it was considered to be a troublesome effect which at times actually interfered with reception. Proposals were advanced from time to time to forbid this method of signaling by international agreement.

The advantages of frequency-shift signaling were recognized as far back as 1923, and several satisfactory systems were demonstrated over the years. It was not until the success of wide-band frequency-modulation transmission in improving signal-to-noise ratio was demonstrated, that narrow-band frequency modulation for telegraph service was recognized as possessing great merit.

Application of this principle resulted in the development of keying involving shifting the carrier frequency alternately between two discrete frequencies, generally separated from 300 to 1000 cycles. One frequency transmits the mark elements and the other the space elements of code.

Receivers employed with this mode of operation are arranged to respond only to frequency variations in the received signals and not to amplitude variations. Inasmuch as the receiving equipment is not affected by large variations in signal amplitude caused by fading, amplitude modulation, etc., and, as noise is predominantly amplitude modulation, frequency-shift operation gives improvements in signal-to-noise ratio averaging from 6 to 12 decibels over on-off single-frequency operation.

Mackay Radio and Telegraph Company has designed special frequency-shift equipment for use on its international circuits. This embodies the principle of space-diversity reception to combat further the effects of signal fading. Profiting by experience under actual operating conditions, further improvements are being incorporated in the equipments, which include transmitter-exciter units, diversity-receiving sets, and monitoring or measuring apparatus.

Many converted radio circuits are now being operated with 5-unit printer apparatus, and time-division-multiplex equipment is being developed. In all of this work, the long experience of Mackay in radiotelegraph operation, combined with that of the associated cable companies in cable and land-wire operation, is providing a wealth of useful information.

Institute of Radio Engineers 1947 Convention

THIRTY-SIX organizations contributed 118 technical papers to the 1947 national convention of the Institute of Radio Engineers in New York, New York, on March 3rd to 6th. Included were 12 from Federal Telecommunication Laboratories, the largest number accepted from any single organization or system. Summaries of these papers follow; certain ones will be published in full in future issues of *Electrical Communication.*

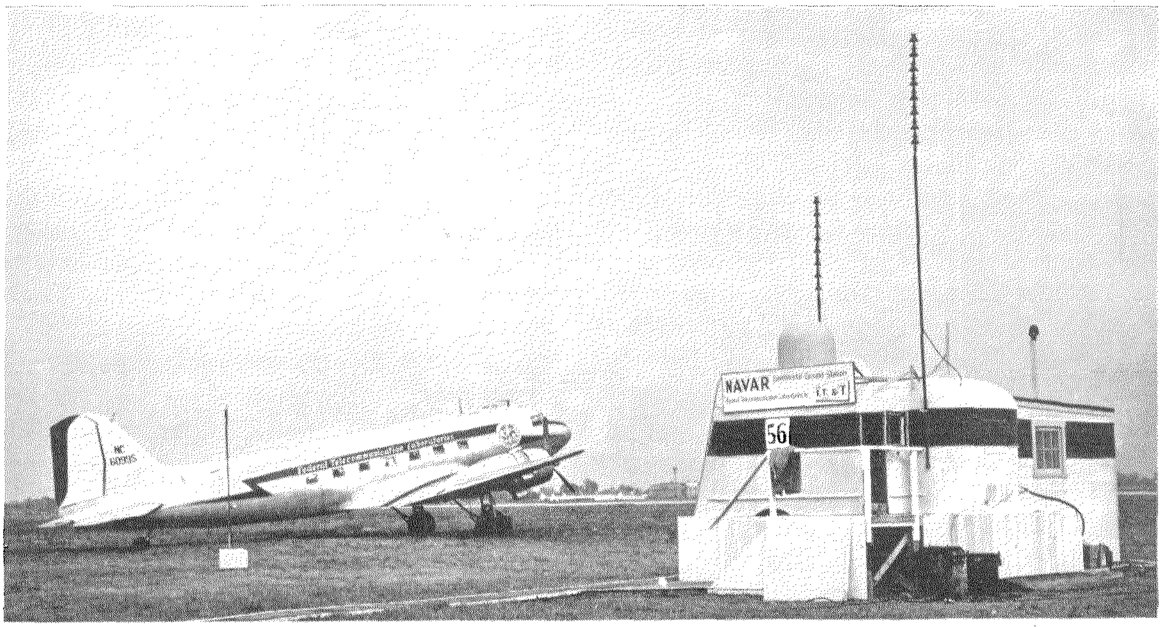
It is of interest to note the effect on this convention of wartime expansion of the telecommunication fields. The attendance was 11,895 and about 170 companies, including Federal Telephone and Radio Corporation, occupied the exhibition space at Grand Central Palace. Banquet facilities restricted attendance at that function to some 1600 members and guests.

First Tests on Navar System for Aerial Navigation and Air Traffic Control

By P. R. ADAMS, S. H. M. DODINGTON, and J. A. HERBST

Background and features of the Navar series of radio aids to air navigation and traffic control, which are being developed under U. S. Army Air Forces sponsorship, are briefly reviewed. A description is given of equipment and its performance in an experimental ground and airplane installation, assembled for preliminary tests of basic principles and the gathering of flight data on the four key facilities of the Navar system, as follows:

A. Airborne distance indicator, pulse type, with meter presentation; B. Airborne azimuth indicator, omnidirectional radio-range facility of physically rotating antenna type, with meter



Navar ground installation and the flying laboratory of Federal Telecommunication Laboratories at Indianapolis for the meeting of the Provisional International Civil Aviation Organization at which the first phase of the Navar development was demonstrated. Two stacks of discone antennas may be seen above the house.

presentation; C. Normal ground search radar, with plan-position presentation; D. Assisted radar, using an airborne responder, with presentation on the same plan-position indicator. Incidentally, signals suitable for operation of a relayed radar plan-position indicator on the airplane were provided.

In particular, features of integrated operation are described: A. Ground radar antenna utilized also as a basis of the azimuth-meter service; B. Ground distance-measuring-equipment responder useful also as an auxiliary transmitter for the azimuth service, as well as for radar relaying; C. Airborne distance-measuring-equipment transmitter performing also as part of radar responder; D. Airborne distance-measuring-equipment receiver, in conjunction with receiver part of radar responder, utilized also for azimuth service and for radar relay service. Station selectivity for azimuth service is attained in spite of common-channel operation of all ground radars. Multipurpose electromechanical type of azimuth-measuring unit is described; principle of time sharing of pulses and method of differentiation between various types of pulse signals on same channel is explained.

Performance of the experimental installation, which was also set up at Indianapolis Airport in October, 1946, for demonstrations to Provisional International Civil Aviation Organization delegates, is discussed. The second stage of development of the system, with emphasis on the earliest availability of a distance-measuring facility, is outlined and features of service-test models of equipment under construction for delivery and operation early in 1947 are described.

Relations Between Bandwidth, Speed of Indication, and Signal-to-Noise Ratio in Radio Navigation and Direction Finding

By H. BUSIGNIES and M. DISHAL

Information is transmitted and received at a very slow rate, electronically speaking, in most navigational and direction-finding systems. Accordingly, quite small bandwidths, probably of the order of only 10 to 100 cycles, are actually required. Thus simple modulation shapes to convey the desired intelligence should be used to conserve bandwidth.

In considering the relations between bandwidth and signal-to-noise ratio, narrowing of the pass band before and after final detection is compared. If large signal-to-noise ratios are always to be used, there is no choice. However, for the important case of small signal-to-noise ratio (e.g., 3 to 1), it is shown that for predetection narrowing, the signal required for a given signal-to-noise ratio is proportional to the square root of the bandwidth ratio, whereas for postdetection narrowing, the signal is proportional to the 4th root of the bandwidth ratio. Experimental confirmation of these relationships is cited.

Impulse noise cannot be limited at a "bandwidth level" equal to that required to pass the signal being received. Illustrations are given showing the excellent results which can be obtained first by impulse limiting in a wide-band portion of the receiver and then passing the signal through the narrow band. For pulse systems using wide bands, a "bucking out" system is described which cancels received impulses but does not affect the signal so long as the impulse and pulsed signal do not occur simultaneously. In this system, the outputs of two diodes are subtracted from each other. Each diode operates from pass bands of equal widths, and the carrier frequency is placed in the middle of only one of the pass bands. Similar results may be obtained by effectively using three pass bands and subtracting the output of two from the third.

A possible new method of reception is described which may allow signals to be reproduced at extremely low signal-to-impulse-noise ratios. This system utilizes the previously mentioned bucking detector scheme described above and measures the resultant impulse-noise output, which is dependent on the signal input, when carrier and impulse are simultaneously present. A graph showing the peak impulse output versus carrier input is given.

The required speeds of indication are considered for various aids to navigation, such as the aircraft radio compass, distance-measuring equipment, long-range navigational systems, Loran, and radar. Considering the rate at which information is transmitted in these systems, it is indicated that in some of the systems apparently more bandwidth is being used than should actually be required. However, it is pointed out that in some of the systems, the extra bandwidth is used to obtain greater accurracies and greater discrimination, e.g., radar.

A practical narrow-band (20-cycle pass bandwidth) navigational system is described. This is the Navaglobe and narrow-band automatic-direction-finding system.

Multiplex Microwave Radio Applied to Telephone Systems

By T. H. CLARK

The first telephone systems used a single wire and a ground return to establish a circuit for a single voice channel of communication. Improved characteristics were obtained when an additional wire was used for the return. It then was found possible to transmit two simultaneous conversations over such a circuit; one between the two wires and ground and the second between the two wires of the pair. The most commonly used long-distance circuits today consist of four wires (two pairs). Three conversations can be transmitted, two on the pairs and a third phantomed between the two pairs. The introduction of carrier techniques and, particularly, the coaxial cable further increased the number of conversations that can be carried over a single pair of conductors. All of these advances were made in the continuous effort to reduce the cost of the wire plant comprised in the telephone system. Despite all these improvements, the wire plant still is the major portion of the cost. The next logical step is the use of radio transmission.

Low-frequency radio was used for long circuits such as transatlantic telephone links at about the time of the first World War. High-frequency circuits for ship-to-shore and intercontinental communication were introduced in the 1920's. Fixed circuits at very-high frequencies came into use in the 1930's for short overwater carries where the cost of the submarine cable was excessive. About the same time, a number of experiments were carried out in ultra- and super-high frequencies for the same service.

At the present time, a number of microwave communication systems, called "microwave links," are under development. The principal purpose is still to reduce the amount of copper used to carry voice-frequency channels. Several of these systems are sufficiently advanced so that microwave links may become a reality in the 1940's as very-high-frequency links did in the 1930's.

Two principal types of systems are in advanced stages in I.T.&T. System laboratories. One system furnishes a single wide-band channel capable of transmitting by radio the frequency spectrum presented by a conventional frequency-division-multiplex carrier system. At the radio receiving terminal, the carrier spectrum is re-obtained and the signal can be applied, as carrier, to wire lines or cables or it can be handled by the receiving terminal of the carrier equipment and be reduced to the separate voice channels.

A second system is capable of taking a number of telephone conversations and applying them as voice bands to time-division-multiplexing equipment. The carrier can be modulated by conventional methods such as frequency modulation, amplitude modulation, or pulse modulation. A system of pulse-time modulation with time-division multiplex has reached a high state of development. A 24-channel system is in experimental operation between New York and Trenton, New Jersey. Two unattended repeaters are used.

Microwaves were exhaustively explored for radar and similar uses during World War II. The techniques evolved have been of considerable assistance in the further development of microwave links. The frequency spectrum above 1000 megacycles is becoming very crowded as a result of the many wartime developments.

Absolute reliability is a primary requisite for telephone systems. The anomalous propagation effects discovered in the early 1930's are being studied more carefully at the present time. Because the microwave systems are limited approximately to line-of-sight distances, repeaters must be used for transmission over long distances. Reliability of these repeaters is of utmost importance and automatic features of protection, switch-over to auxiliary equipment or power, and operation under extremes of temperature and humidity have been carefully studied. Repeaters, to be economical, must be capable of reliable unattended operation.

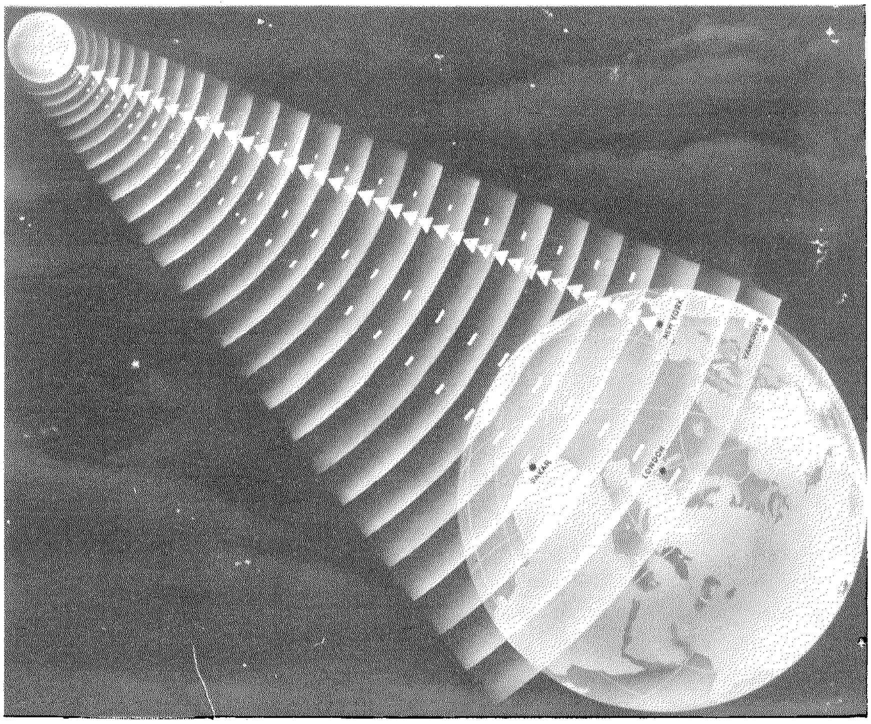
Telephone systems have attained a perfection and a complexity that is awe inspiring. The radio engineer must engineer the radio system to have the same reliability and to accept any of the complex systems of dialing, signaling, supervision, etc., which are in common use in the telephone plant.

Considerations of Moon Relay Communications

By D. D. GRIEG, S. METZGER, and R. WAER

The moon offers possibilities as a passive repeater for radio links. Such a system requires the evaluation of various phenomena, which are of no importance in conventional radio communication systems. One such phenomenon is the Doppler shift in the frequency of the received signal as a result of the relative motion of the earth and the moon. Another is possible interference by electromagnetic disturbances from the sun or other astral bodies, i.e., cosmic noise. It must also be remembered that contact can be established only while the moon is visible at both the transmitting and receiving points.

As the exact nature of the radio reflecting properties of the surface of the moon is not known, calculations have been made for two extreme cases. The first assumes the moon to be perfectly smooth, while the second assumes the moon to be a perfectly diffuse reflector. The first assumption indicates that signals of all types, including telegraph, speech, and even color television, would be reflected without distortion. On the other hand, the second assumption leads to the conclusion that the distortions are such that the transmission of telegraph or teletype signals is practicable, narrow-band speech is possible, but the transmission of television signals is doubtful.



Hemispheric broadcasting, utilizing the reflecting properties of the moon, offers interesting possibilities. Radio transmissions beamed at the moon are reflected back to earth and irradiate all points on earth facing the moon.

The conclusions from either assumption indicate that the transmitting power required for communication by telegraph, teletype, or speech, is available with existing equipment. However, the requirements for large antennas and facilities for continuously tracking the moon seem beyond the reach of a home receiver for this purpose, so that for the time being moon-reflected signals would have to be utilized by rebroadcasting.

The meager information available from the United States Signal Corps moon radar experiments indicates that the amplitude of successive echoes varies widely. An attempt has been made to explain this phenomenon by assuming the surface of the moon to be diffuse, but with a number of smooth spots, each not necessarily more than two or three miles in diameter. The echoes from smooth spots of this size have been shown to be alike and about the same as that from the whole diffuse moon. The echoes resulting from a surface of the type assumed thus could interfere with each other in such a manner as to add to one pulse and cancel a pulse sent several seconds later. Calculations indicate that two such spots located most anywhere on the moon would produce the results observed in the moon radar experiments.

Power Loads at Very- and Ultra-High Frequencies

By A. G. KANDOIAN and R. A. FELSENHELD

The design of satisfactory loads to handle and measure large amounts of power at very- and ultra-high frequencies has always been an annoying, though not necessarily a difficult, problem. There are a variety of possible solutions but each has its own particular defects. Some have poor impedance characteristics and are nonlinear, some cannot stand large voltage or heat gradients, some are very difficult to calibrate accurately, and others dissipate an appreciable portion of the power in electromagnetic radiation. Nearly all are difficult to cool if large powers are involved.

To overcome some of these defects, a study has been made of several new designs, which appear to have definite advantages over existing units.

TRANSMISSION-LINE LOAD

The first of these is a transmission-line type of load. It is made up of a length of coaxial line using water as dielectric between the center and the outer conductor. The water is circulated to provide necessary cooling. Because of the high dielectric constant of water, approximately 80, it is not convenient to obtain a desired surge impedance in the neighborhood of 50 ohms. A surge impedance of 20 ohms, however, is easily feasible with a nominal-sized outer conductor. As a result of high dielectric constant and high propagation loss in water, an equivalent infinite line is easily obtained with a total length in air no greater than one wavelength. Transformation from 20 to 50 ohms is readily accomplished by one or more quarter-wave transformers.

In practice, such a load has been used for average powers of the order of a kilowatt and peak powers of several hundred kilowatts in the neighborhood of 500 to 1000 megacycles. It is adaptable to both lower and higher frequencies, and for higher average power.

RADIATOR-TYPE LOAD

Another load, which has proved useful and very convenient, is of the radiator type. The radiator or antenna is totally enclosed in a tank of water. Again, as a result of the high dielectric constant and high losses, the radiated electromagnetic energy is converted to heat in the water, which may be circulated for cooling purposes. The amount of power is easily calculated from the rate of flow and the difference in temperature between input and output water.

In practice, such a load has been used for average powers of the order of one kilowatt and peak powers of several hundred kilowatts. It is, however, not quite as satisfactory from a peak-power standpoint as the transmission type of load.

RESONANT-CAVITY TYPE OF LOAD

A third type of load is the resonant-cavity type. It is well known that one may couple into a tuned cavity so that the input impedance will be any desired value, a particular example being a 50-ohm pure resistance. If the Q of the cavity is very high, the 50-ohm relationship will hold over a very narrow frequency range. Under these conditions, any amount of power transmitted to this load will be dissipated in the ohmic loss within the cavity. Because of the high Q of the cavity, very large circulating currents will be set up along with extremely high voltages in the current-minimum region.

When such a cavity is specifically designed as a load, however, the Q is deliberately made low. By using a quarter-wave coaxial resonant circuit and Kovar tubing for the inner conductor, Q's of the order of 30 may be obtained. The coupling consists of a direct tap into the center conductor near the current-maximum point.

Cooling is obtained by circulating water through the Kovar tubing which forms the center conductor, where the dissipation is concentrated.

A load of the resonant-cavity type has been used for testing frequency-modulated broadcast transmitters. Average powers of the order of 10 kilowatts have been handled and such a design for 50 kilowatts appears feasible.

Ultra-High-Frequency Multiplex Broadcasting System

By A. G. KANDOIAN and A. M. LEVINE

Because of the apparent need in the near future for a supplement to amplitude- and frequency-modulation broadcasting, time-division multiplexing and pulse-time modulation has been developed. This ultra-high-frequency system may be used for high-fidelity low-noise broadcasting for educational and other purposes.

Pulse-time modulation was chosen because of the many advantages it offers at ultra-high frequencies over other systems in size, simplicity of equipment, economy of power, high-quality low-noise reception, and improved stability. High transmitting efficiency is obtained with pulse-time modulation because the signal is converted to a sequence of pulses of very short time duration. The transmitter is, therefore, inoperative most of the time and a very low average power produces high peak powers when the transmitter is keyed. A further improvement is realized by the use of a high-gain omnidirectional transmitting antenna, which concentrates the power radiated vertically in a thin fan-shaped beam extending to the horizon.

At ultra-high frequencies, coverage area is limited approximately to line-of-sight distances. Maximum coverage for the service area is assured by centrally locating the transmitter at the highest elevation possible.

As all programs originate from the same transmitter, a highly directive receiving antenna is utilized which is permanently focussed on the transmitter. This feature reduces interference from near-by buildings or other fixed objects.

The receiver itself is fixed tuned to one transmission frequency, eliminating the cost of complex ganged tuning controls. Program selection may be accomplished by push-button selection of simple timing circuits. This makes the problems of oscillator drift and warm-up period unimportant.

The particular pulse-time modulation system that has been developed simultaneously transmits eight programs of 9500 cycles each over a single radio-frequency carrier of 930 megacycles.

The eight programs pass through control panels to the time-modulation mixer. Here the audio-frequencies are converted to 0.5-microsecond pulses at a base repetition rate of 24,000 per second. The various channels are then mixed and a marker added. The video-frequency output consists of a series of pulses having a bandwidth of 3 megacycles. A 70-ohm coaxial cable transmits the pulses to the modulator.

The receiving system consists of a directional antenna, a fixed-tuned radio-frequency section, and a pulse-time demodulator. The radio-frequency section includes a local oscillator, crystal mixer, and a 30-megacycle intermediate-frequency preamplifier. The pulse-time demodulator includes final intermediate-frequency amplification, video-frequency detection and tuning, and audio-frequency detection and amplification.

New home of the Institute of Radio Engineers on Fifth Avenue in New York City. Built about 1880, this replica of a portion of Chemonceaux, a famous French chateau, was one of the prominent mansions in an era of elegance. Funds for the purchase, refurbishing, and furnishing of the building were contributed by Institute members and the radio industry.

Broad-Band Very-High-Frequency Amplifiers

By A. M. LEVINE and M. G. HOLLABAUGH

The present trend in ultra-high-frequency receivers is toward use of increasingly higher intermediate frequencies. Broad-band amplifiers are being constructed for operation in the upper part of the very-high-frequency range. This is a region of transition from lumped to distributed circuit parameters, where tube damping, undesired feedback, and related effects assume important magnitudes. These effects, in fact, exercise a controlling influence in many designs.

Calculations for input damping, instability caused by feedback, and similar factors have been made. Calculated and measured values are given for various tube types and for frequencies between 30 and 300 megacycles.

Graphed values of effective input resistance were found to follow the customary frequency-squared approximation in the lower very-high-frequency region. For higher frequencies, smaller resistances were observed. For design purposes, $f^{2.5}$ lines have been found to be sufficiently accurate with three common tube types, 6AG5, 6AH6, and 6AK5.

The tube damping values were taken for amplifier stages where short cathode leads were used but without other means of input-admittance compensation. The input resistances determined under this condition were considerably smaller than those measured by a Q-meter with all electrodes except the grid at ground potential. For example, the Q-meter input resistance of a 6AK5 at 100 megacycles was in the 15,000-ohm region, whereas the resistance obtained from the same tube in the amplifier stage was in the vicinity of 5000 ohms.

It was shown that tube damping limits the minimum bandwidth obtainable, unless special precautions are taken to eliminate the effect. This minimum bandwidth may be expressed by equations or curves.

Tube damping must also be considered in connection with the choice of interstage coupling. The more complex interstage networks, i.e., those designed for greater gain and bandwidth capabilities, require higher termination or damping resistances and this may easily be in excess of the tube damping resistance present.

It is generally recognized that at high frequencies it is difficult to obtain large stable gains when the bandwidth is very small relative to the carrier frequency. Intermediate-frequency amplifiers for various frequency-modulation receivers often present this problem. It is shown that capacitive coupling caused by the grid-to-plate capacitance in amplifier tubes is sufficient to give rise to instability in amplifiers in the very-high-frequency region. A limit to the minimum bandwidth obtainable with a given gain is determined as a function of the grid-plate capacitance of the amplifier tube. Or conversely, a limit to the maximum gain obtainable with a given bandwidth exists as a function of the grid-plate capacitance. A 100-megacycle amplifier, using four stages of 6AK5's, will be unstable if the bandwidth is less than approximately 2 megacycles, assuming that the gain is not intentionally reduced or that some form of neutralization is not used.

Compensation of Phase Shift at Low Frequencies

By F. McGEE

The formerly unimportant low-frequency phase shift has recently come into the position of a first-rate problem. Terman, Hewlett, Palmer, and Pan,[1] and Sturley[2] have analyzed the resistance-capacitance amplifier and presented formulas and graphs representing low-frequency phase shifts in terms of circuit elements and tube constants. These relationships may be presented in somewhat more practical form for design purposes. A new method of phase compensation, which allows for the simultaneous correction of phase errors in cathode, screen, and coupling networks, is discussed.

The basic circuit equations

$$\left.\begin{aligned} i_p &= g_m e_g + g_s e_s \\ i_s &= G_m e_g + G_s e_s \\ e_g &= e - e_k \\ e_s &= -i_s Z_s - e_k \\ e_k &= (i_p + i_s) Z_k \end{aligned}\right\} \quad (1)$$

in which:

i_p = plate alternating current
i_s = screen alternating current
e_g = grid-to-cathode alternating voltage
e_k = cathode-to-ground alternating voltage
e = grid-to-ground alternating voltage
Z_s = screen-to-ground alternating-current impedance
Z_k = cathode-to-ground alternating-current impedance
g_m = grid-to-plate transconductance
G_m = grid-to-screen transconductance
g_s = screen-to-plate transconductance
G_s = screen-to-screen conductance (reciprocal of the dynamic screen resistance r_s)

are solved simultaneously by means of fifth-order determination yielding:

$$i_p = \frac{g_m e}{1 + G_s Z_s + G_k Z_k}, \quad (2)$$

in which for simplicity

$$G_k = g_m + g_s + G_m + G_s. \quad (3)$$

Assume for the moment that the screen impedance to ground is zero. By algebraic manipulation, the approximate formula

$$C_k = \frac{9 G_k}{\phi_k f} \quad (4)$$

where ϕ_k is the phase shift in the cathode circuit at the frequency f caused by the capacitor C_k providing the phase shift is less than 10 degrees, and G_k and R_k have normal values. Then assuming that the cathode phase shift is negligible and using similar manipulation and conditions,

$$C_s = \frac{9}{\phi_s f r_s}, \quad (5)$$

where ϕ_s is the phase shift caused by the screen by-pass capacitor C_s at the frequency f when the dynamic screen resistance is r_s. It is then shown that under the condition named, the total phase shift, when both cathode and screen contribute, is the sum of ϕ_k and ϕ_s as given by these two formulas. It is observed that any phase shift resulting from the impedance in the plate circuit adds to the phase error in the current as does that caused by a coupling network. Under video-frequency conditions, the phase shift resulting from plate decoupling is derived and, by algebraic manipulation, is put in the form

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[1] F. E. Terman, W. R. Hewlett, C. W. Palmer and W. Y. Pan, "Calculation and Design of Resistance-Coupled Amplifiers Using Pentode Tubes," *Transactions of the A.I.E.E.*, v. 59, p. 879; 1946.

[2] K. R. Sturley, "Low Frequency Amplification," *Electronic Engineering*, November and December, 1944; and January, February, March, April, and May, 1945.

$$C_p = -\frac{9}{\phi_p f R_L}, \tag{6}$$

where the phase shift ϕ_p is caused by the decoupling capacitor C_p at the frequency f, and R_L is the plate load resistor. Again ϕ_p must be less than 10 degrees and R_L must be at least equal to the decoupling resistor. The phase shift in the coupling network is derived and put in the form

$$C_c = \frac{9}{\phi_c f R_c}, \tag{7}$$

where the capacitor C_c causes the phase error ϕ_c at the frequency f. R_c is the output resistor and is assumed large with respect to the impedance in the preceding plate circuit.

It is observed that if all conditions are met simultaneously, the total phase shift ϕ_T is

$$\phi_T = \frac{9}{f}\left(\frac{G_K}{C_k} + \frac{1}{C_s r_s} + \frac{1}{C_c R_c} - \frac{1}{C_p R_L}\right). \tag{8}$$

It is, therefore, possible to obtain zero phase shift over the stage by making

$$\frac{G_k}{C_k} + \frac{1}{C_s r_s} + \frac{1}{C_c R_c} = \frac{1}{C_p R_L}. \tag{9}$$

Noise-Suppression Characteristics of Pulse Modulation

By S. MOSKOWITZ and D. D. GRIEG

The utilization of pulses for the multiplexing of communication channels has received considerable attention during the past few years. In an analogous manner to the well-known frequency or phase modulation of continuous waves, pulse-time modulation offers an improvement in signal-to-noise ratio over that obtained by the common amplitude modulation of either continuous or pulsed waves. It is shown that the improvement is proportional to the radio-frequence bandwidth used in the transmission link. In terms of the pulses, the improvement is proportional to the time-modulation displacement and inversely proportional to the build-up or decay time, whichever is the smaller.

An important measure of protection against noise interference offered by time-modulated pulses results from the high ratio of peak-to-average power used. The threshold of improvement is reached when the peak pulse amplitude is about twice the effective noise peaks. Hence, devices such as limiters may be used to considerable advantage. Experiments have been made and the results are given illustrating the outlined theory. The effectiveness of various noise-suppression devices such as limiters, differentiators, and multivibrators is demonstrated by experimental data.

The greatest degree of noise suppression is obtained when successive stages of limiting and differentiation are incorporated in the receiver. This result may be understood by considering the following manners in which noise can enter the pulse system:

A. Amplitude modulation of the pulses.
B. Width modulation of the pulses.
C. Noise occurrence between pulses.
D. Displacement in time of the leading or trailing edge of the pulses.

Noise entering by amplitude modulation of the pulses and between the pulses may be removed by proper limiting providing the input signal-to-noise ratio is greater than 6 decibels. Following this stage, a differentiator serves to extract the proper pulse edge thus removing width-modulation noise. However, as some edge-slope variation may be introduced by noise, the output of the differentiator may again contain some amplitude and width noise. Such secondary noise modulations may be further suppressed by successive stages of limiting and differentiation. The operations may also be obtained by the action of a multivibrator.

Noise entering by displacement in time of the leading or trailing edge of the pulse is of the same form as the modulating signal and is inherent in the system of modulation. However, the noise displacement may be reduced by decreasing the build-up or decay time of the pulses, i.e., increasing the bandwidth of the system.

In a similar manner, impulse noise, such as that derived from electrical machinery, automobile ignition, and interfering pulse communication systems, may be suppressed. The degree of suppression is usually greater than that obtained with random noise.

Video-Frequency Negative-Feedback Amplifiers

By M. G. HOLLABAUGH, J. A. RADO, and A. M. LEVINE

Video-frequency amplifiers having bandwidths of four or five megacycles may be readily designed using well-known methods such as shunt, series, or series-shunt peaking to achieve a uniform gain over the required frequency band. These techniques have been developed largely empirically and are adequate for relatively narrow bands and low gains. Design alignment is simple and the technique is widely known.

In color-television and pulse-modulation systems, the need for much wider bands in the range from 10 to 50 megacycles is encountered and it is necessary to exploit as fully as possible the capabilities of components to realize a practical result. Almost ten years ago, H. A. Wheeler foresaw this need.[1] His paper set down the fundamental relation that the gain-bandwidth product equals the transconductance over the geometric mean of the input and output capacitances. Filter theory was applied to the problem of interstage coupling of wide-band amplifiers and resulted in the realization of as high as 95 percent of the above criterion in practical amplifiers.

The complexity of design and manipulation of these high-performance filters, each of which may require adjustment of five or six interdependent parameters, led to the consideration of a simpler means of attaining equivalent performance, that of inverse feed-back. Wheeler proposed the use of inverse feed-back but the development and use of it was largely limited to band-pass or intermediate-frequency amplifiers and was shrouded by wartime security regulations.

Simplicity is the most important advantage to be gained in using feedback and the advantage increases with the number of amplifier stages for two reasons. First, adjustments need be made only on the terminating filter of the whole amplifier instead of on each interstage coupling filter because the interstage coupling consists of only a feedback resistor. Second, the performance of a feedback amplifier approaches more closely the theoretical criterion as the number of stages is increased. Other advantages arise out of the use of feedback, such as the reduction of distortion resulting from nonlinearity.

Analysis of the generalized feedback amplifier reveals its similarity to a transmission line or ladder network. In fact, it is a ladder network with negative-conductance shunt arms. This is the function of the amplifier transconductance which is a negative quantity. A mathematical analysis has been based on this assumption. The solution for stage gain established the fact that these amplifiers have a gain-bandwidth capability equal to that of an ideal amplifier. In actual amplifiers, the ideal has been approached very closely.

Using 6AK5 tubes, amplifiers have been constructed having video-frequency bandwidths of 13 megacycles and 20 megacycles. In these amplifiers, simple terminations were used consisting of the characteristic resistance of the network in series with a peaking inductor.

A means has been developed for correcting the low-frequency response of the amplifier resulting from the practical limitation on the size of the interstage coupling capacitor. As this requires the introduction of shunt resistance, an examination was made of the effect of shunt dissipation. It was proved that shunt resistance may be reduced to a value as low as that of the feedback resistance, at which point for equal bandwidth the gain per stage is reduced by ½ decibel. Larger values have negligible effect.

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[1] H. A. Wheeler, "Wide-Band Amplifiers for Television," *Proceedings of the I.R.E.*, v. 27, pp. 429–437; July, 1939.

Ideally, it has been pointed out that the feedback path should be unidirectional in the backward direction just as G_m is unidirectional in the forward direction. Actually, if the network is properly terminated, this is unimportant as the mathematical analysis shows the backward traveling wave is negligible and decreases.

Circularly Polarized Antennas

By W. SICHAK and S. MILAZZO

A formula is derived which gives the variation in received voltage as an elliptically polarized antenna is rotated in a plane transverse to the direction of propagation of an incident elliptically polarized wave. It is shown that a circularly polarized antenna will not receive any reflected energy from a smooth metallic reflector while an elliptically polarized antenna will receive some energy, the magnitude of which depends on the ratio of the axes of the polarization ellipse. If circular polarization is used for communication, both the transmitting and receiving antennas must produce the same screw sense of polarization, otherwise no signal will be received. If the antennas are elliptically polarized, some signal will always be obtained. Some methods for obtaining circular polarization are discussed. For the case of a horizontal loop and vertical dipole, the currents in the loop and dipole must be in phase to get circular polarization. Experimental results are given to confirm the theory.

Monitoring Equipment for Frequency-Modulation Broadcasting

By M. SILVER

This paper deals with the design of monitoring equipment for frequency-modulation broadcasting and in the measurements required to prove performance.

Briefly, the equipment operates as follows: A portion of the transmitter output is mixed with a crystal standard so that the difference frequency is 150 kilocycles. This voltage is applied to a counter-type discriminator which is used both to measure the line transmitter frequency and monitor its audio-frequency output. The required linearity for this application is extremely high and in the case of this equipment it is 0.02 percent. The output of the counter discriminator is filtered and then applied to an audio-frequency amplifier for acoustic monitoring and measurement purposes. In addition, the discriminator feeds a suitable vacuum-tube voltmeter for percentage-modulation measurements and a thyratron flasher circuit for overmodulation indication.

The equipment is capable of meeting the requirements of the Federal Communications Commission and surpasses them in many aspects. The monitor is capable of measuring noise to -80 decibels and distortion of 0.2 percent. Station carrier frequency can be measured to an accuracy of ± 100 cycles under full-modulation conditions with a long-time stability of one part in 5×10^5.

Recent Telecommunication Developments

McVey Awarded Kernot Medal—Daniel McVey, Chairman of the Board and Managing Director of Standard Telephones and Cables Pty Ltd., has been awarded the Kernot Memorial Medal.

This medal was established to perpetuate the memory of William Charles Kernot, the first professor of engineering at the University of Melbourne. It is awarded

> For distinguished engineering achievement in Australia and for that achievement which is placed first in order of merit by the Assessors.

The three assessors, who are appointed by the council of the university, were unanimous in designating Mr. McVey as the recipient of the medal.

* * *

Etchings of Heinrich Hertz—Starting in 1935, a series of etchings of men famous in the telecommunication field has been issued by Bureau de l'Union Internationale des Télécommunications, Effingerstrasse 1, Berne, Switzerland. Recently added to the series is an etching of Heinrich Hertz. Reproduced on luxury paper, each portrait measures 23 by 17 centimeters (9 by 6⅝ inches) including margins. Copies of etchings of Hertz, Morse, Hughes, Bell, Marconi, Baudot, Gauss and Weber, Maxwell, Ferrié, Siemens, Popov, and Ampere may be obtained for 3 Swiss francs each, including carriage and packing. Hertz is credited with having discovered the wave character of electrical transmission through space and through wires. His work on wave propagation paved the way for the development of wireless telegraphy. Always an exacting scientist, Hertz demonstrated his principles with carefully worked out experiments. These demonstrations proved that electromagnetic waves, or Hertzian waves as they became known, could not only be transmitted through space, but could be reflected, absorbed, polarized, and made to perform in other ways similar to light waves.

* * *

Miniature Thermal Delay Switch—Standard Telephones and Cables (London) has developed a delay switch of the thermal type which has been coded ULS631. It may be used to control the time between applying heater voltage and anode voltage in indirectly heated vacuum tubes or gas-filled rectifiers. The elements are mounted in an evacuated glass envelope having a seated height of 1¾ inches and a diameter of ¾ inch. Its ratings are as indicated in the table below.

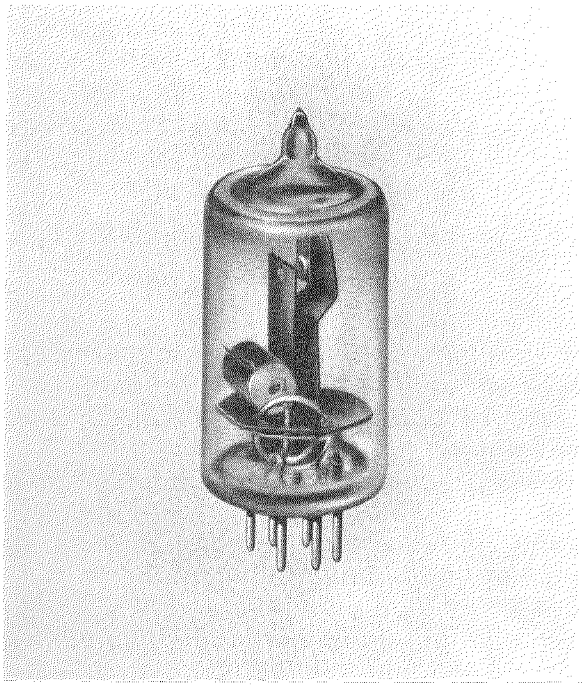

Table of Characteristics

Heater, Volts	6.3
Nominal Heater Current, Amperes	0.5
Nominal Delay at 20 Degrees Centigrade, Seconds	50–60
Ambient Temperature Range, Degrees Centigrade	−350 to +85
Maximum Open-Circuit Voltage Between Contacts, Direct Volts	220
Maximum Contact Current on Make, Amperes	1.0
Maximum Surge Current on Make, Amperes	5.0
Maximum Contact Current on Break, Milliamperes at 50 Direct Volts	100

Contributors to This Issue

TREVOR H. CLARK

TREVOR H. CLARK was born in Kansas on July 16, 1909. He received a B.S. degree in physics and mathematics from Friends University in Wichita, Kansas, in 1930, and an M.S. degree in physics from the University of Michigan in 1933. He was associated with the R.C.A. Manufacturing Corporation until 1938, when he joined the research laboratories of International Telephone and Telegraph Corporation.

Until 1940, he was engaged in research on vacuum tubes in the Paris laboratories of the I.T.&T. System. He returned to the United States in 1941, after a survey trip in Portugal and Portuguese West Africa, and assisted in the organization of the New York laboratories, which later became Federal Telecommunication Laboratories. During the war, he conducted research in direction finding, primarily for the United States Navy. In 1945, he began work with pulse-time-modulation relay systems. He has recently been appointed Manager of Technical Services of Federal Telecommunication Laboratories, Inc.

Mr. Clark is a Senior Member of the Institute of Radio Engineers, and a member of the Acoustical Society of America.

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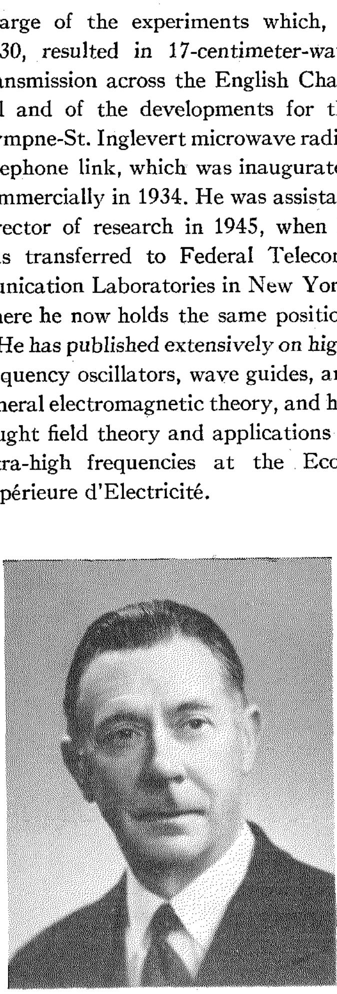
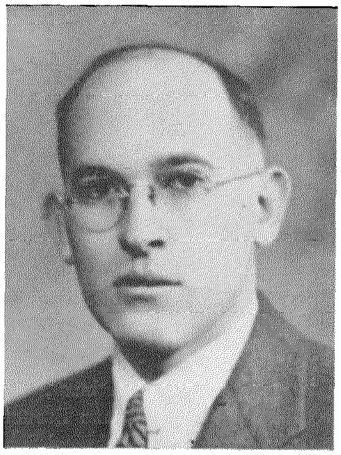
A. G. CLAVIER was born in Cambrai, France, in 1894. He received a degree in electrical engineering from Ecole Supérieure d'Electricité in 1919 and then joined the staff of engineers organized by General Ferrié at the Etablissement Central de la Radiotélégraphie Militaire. He was in charge of research on high frequencies from 1920 to 1925.

In 1929, Mr. Clavier joined Les Laboratoires Standards in Paris which later became Laboratoire Central de Télécommunications, and has been continuously engaged in research on centimeter and millimeter waves. He was in charge of the experiments which, in 1930, resulted in 17-centimeter-wave transmission across the English Channel and of the developments for the Lympne-St. Inglevert microwave radiotelephone link, which was inaugurated commercially in 1934. He was assistant director of research in 1945, when he was transferred to Federal Telecommunication Laboratories in New York, where he now holds the same position.

He has published extensively on high-frequency oscillators, wave guides, and general electromagnetic theory, and has taught field theory and applications of ultra-high frequencies at the Ecole Supérieure d'Electricité.



A. G. CLAVIER



ROBERT I. COLIN

Mr. Clavier is president of the section of the Société des Radioélectriciens dealing with hyperfrequencies. He is a Fellow of the Institute of Radio Engineers, and a Member of the Institution of Electrical Engineers.

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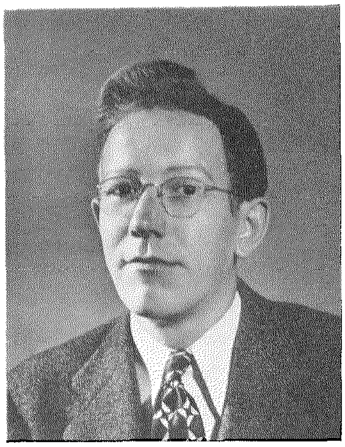
ROBERT I. COLIN was born in Brooklyn, New York, on Feb. 16, 1907. He attended Cornell University as a student in engineering and physics under two New York State scholarships. During his senior year he was on the physics staff as an undergraduate assistant. He received the A.B. degree in 1928, and was elected to the honorary scholastic society Phi Kappa Phi.

The next year, under an exchange fellowship of the Institute of International Education, he attended the University of Frankfurt A/M, Germany, majoring in physics; and visited many European countries. From 1929 to 1933, he was a graduate assistant on the physics staff of New York University, teaching and taking graduate courses. He left with an M.S. degree after doing an experimental thesis in the field of electron-optics. From 1934 to 1939, he was instructor in physics and related subjects at the Hebrew Technical Institute. From 1941 to 1944



P. I. CORBELL, JR.

he was instructor, then head, of the Aircraft Electrical Systems Branch of the Army Air Forces Technical School, first at Chanute Field, Illinois, and then at Yale University.

Mr. Colin entered the Federal Telecommunication Laboratories in September, 1944, as a technical writer, then senior engineer, where he did technical studies, wrote proposals, and shared in the development of a number of proposals in the field of radio air navigational aids, especially for long distances, including the "Navaglobe" system.

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P. I. CORBELL, JR. was born at Napa, California, in 1919. He received a B.S. degree in electrical engineering from the University of California in 1943. During his attendance there, he was employed by Pacific Gas and Electric Company, San Francisco, as an engineering assistant.

In 1943, he joined the vacuum-tube department of Federal Telephone and Radio Corporation, where he was engaged in transmitting-tube development. Since November, 1946, he has been with Litton Engineering Laboratories, Redwood City, California, working on the development of continuous-wave magnetrons.

SIDNEY FRANKEL

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SIDNEY FRANKEL was born on October 6, 1910 in New York City. Rensselaer Polytechnic Institute conferred three degrees on him: the B.A. degree in electrical engineering in 1931, and in mathematics the M.A. degree in 1934 and the Ph.D. degree in 1936. He was an instructor in mathematics from 1931 to 1933.

For a year after leaving college, Dr. Frankel served as a sound-recording engineer with the Brooklyn Vitaphone Corporation. In 1937–1938, he was an assistant engineer in the design and development of electronic flight instruments for the Eclipse Aviation Corporation.

He joined the Federal Telegraph Company staff at Newark, New Jersey, in 1938 as an engineer on the design and development of radio transmitters. In 1943, he was transferred to Federal Telephone and Radio Laboratories, now Federal Telecommunication Laboratories. At present he is engaged in the development of components for microwave systems.

He is a member of Sigma Xi and a Senior Member of the Institute of Radio Engineers.

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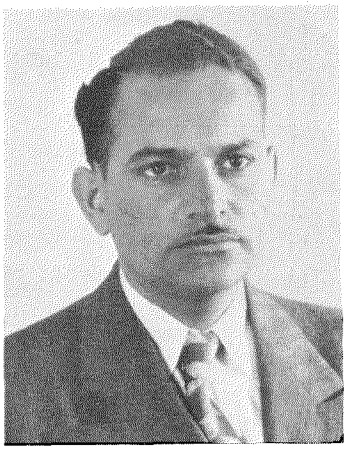

HARRIS GALLAY

HARRIS GALLAY was born in New York City on July 28, 1908. He received his B.S. degree from the College of the City of New York in 1931. From 1935 to 1943, he was associated with the U. S. Army Signal Corps, as a civilian radio engineer in the development and design of instrument landing and airborne radio compass equipment, and in aircraft radio maintenance and repair facilities.

He joined the engineering staff of Federal Telecommunication Laboratories in 1943, and is now engaged in supervising microwave communication and control-equipment projects utilizing pulse-time modulation.

Mr. Gallay holds an Associate membership in the Institute of Radio Engineers.

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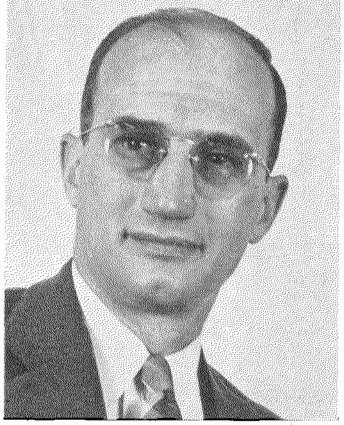
JOHN J. GLAUBER was born in New York, New York, on July 31, 1903. He received the M.E. degree from Stevens Institute of Technology in 1925.

In 1925, he obtained a position with the U.S. Tool Company in Ampere, New Jersey, on variable-capacitor design. In 1927, he joined the Arcturus Radio Tube Company, Newark, New Jersey, as laboratory assistant and was chief engineer from 1933 to 1936. He then joined the Westinghouse Lamp



JOHN J. GLAUBER

George H. Gray

Company, Bloomfield, New Jersey, as a vacuum-tube development engineer and in 1939 became development engineer for the National Union Radio Corporation, Newark, N. J. He was employed by the Clark Controller Company, Cleveland, Ohio as development engineer on special gas-filled control tubes, and by the Champion Radio Works, Danvers, Massachusetts, as quality engineer.

He joined the vacuum-tube department of Federal Telecommunication Laboratories, New York, in 1941, becoming department head in charge of the design and development of tubes. In 1947, Mr. Glauber joined the engineering staff of United Electronics Corporation of Newark, New Jersey.

Mr. Glauber is a Senior Member of the Institute of Radio Engineers.

* * *

D. D. Grieg

George H. Gray was born in Arlington, Massachusetts, on October 8, 1885. He received an S.B. degree from Massachusetts Institute of Technology in 1909 and an E.E. degree from the University of Wisconsin in 1916.

He served as an instructor at Massachusetts Institute of Technology from 1909 to 1910. After two years with the Telluride Power Company, he spent a year with the Western Electric Company. From 1913 to 1917, he instructed in electrical engineering at the University of Wisconsin. He then went to the Northwestern Bell Telephone Company for nine years.

In 1926, Mr. Gray joined the International Telephone and Telegraph System where he served with the parent corporation and International Standard Electric Corporation except for one year (1940–1941) with the U. S. Signal Corps.

He is a Member of the American Institute of Electrical Engineers.

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D. D. Grieg was born on February 26, 1915 in London, England. He received his early schooling in England and the B.S. degree in electrical engineering from the College of the City of New York.

From 1936 to 1940, he was in charge of the television department of the Davega Radio Company. In early 1941 he taught radio communication in the Brooklyn Technical High School. Since 1941, he has been a research engineer for Federal Telecommunication Laboratories. He is now a Division Head and has charge of the Television and Communication Departments.

Mr. Grieg is a Senior Member of the Institute of Radio Engineers and a Member of the American Institute of Electrical Engineers. He has served on several technical committees including the Television Committee of the Radio Technical Planning Board and those on Television Relays and Studio-Transmitter Links of the Radio Manufacturers Association. He is the author of several technical papers and holds many patents in the field of radio.

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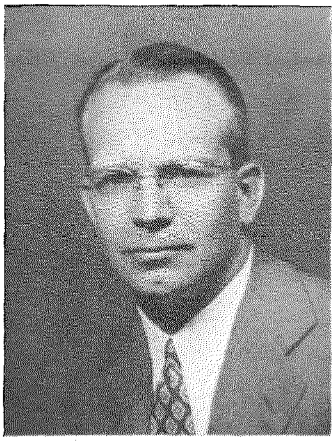

Ross B. Hoffman

Ross B. Hoffman was born in St. Louis, Missouri, on September 30, 1907. He received the B.S. degree in electrical engineering from the University of Illinois in 1929, and joined the I.T. & T. system on graduation. He spent five years in Buenos Aires on installation and operation engineering of transatlantic radiotelephone and radiotelegraph equipment for the Compañía Internacional de Radio Argentina. From 1935 to 1938, he was a development engineer in the Les Laboratoires, Le Matériel Téléphonique in Paris, France.

After returning to the United States in 1938, he was associated with International Telephone Development Corporation and Federal Telegraph Company on the development of instrument landing systems until January 1941, when he was transferred to the new Federal Telephone and Radio Laboratories in New York.

August F. Jones

EMILE LABIN

Late in 1943, he transferred to the Federal Telephone and Radio Corporation in Newark, and since that time has been engaged in development and product engineering in connection with aerial navigation and, more recently, on equipment for mobile communication.

• • •

AUGUST F. JONES received a B.S. degree from Denison University in 1922 and an E.E. degree from Cornell University in 1925.

He was then employed by the International Telephone and Telegraph Corporation as a student engineer. After a year in the equipment engineering department of the Cuban Telephone Company, he was transferred to Spain to organize a commercial department in the northwest district of the Spanish National Telephone Company. In 1928, he returned to New York where he spent three years in the engineering department, after which he was transferred to the manufacturing division of the company. He was assigned to the Madrid factory in 1933, and was made chief engineer in 1939. In 1942, he returned to the U.S.A. and is at present in the telephone engineering department of Federal Telephone and Radio Corporation.

ELLIOTT W. MARKOW

• • •

EMILE LABIN was born on December 1, 1907 in Bucharest, Rumania. He graduated from Ecole Polytechnique in Paris in 1930.

On finishing school, he become associated with Television Baird-Natan, the Paris associate of Baird Television Ltd. of England. In 1932, he became a television research engineer for Compagnie de Compteurs. He joined the engineering staff of Le Matériel Téléphonique in 1935, where he developed the Eiffel Tower television station, then the most powerful in the world, in addition to work on transmitters, vacuum tubes, and radar.

ERNEST S. MCLARN

In 1940, Mr. Labin was transferred to the U.S.A. to take charge of the division which later became Federal Telecommunication Laboratories.

He is the author of numerous papers on television and radio published in both French and U.S.A. technical journals. Mr. Labin is a Senior Member of the Institute of Radio Engineers.

After the war, Mr. Labin was appointed techinal director of International Telecommunication Laboratories in charge of coordination of research and development activities of I.T.&T. System Associated Companies.

GEORGES PHELIZON

• • •

ELLIOTT W. MARKOW was born in St. Louis, Missouri, on April 2, 1921. He obtained a B.S. degree in electrical engineering from Purdue University in 1942. Directly after graduation he joined Federal Telephone and Radio Corporation.

He worked on production and development of various types of military equipment during the war years. His present activities are in the application of mobile equipment to radiotelephone services.

• • •

ERNEST S. MCLARN was born in Oakdale, Pennsylvania, on November 30, 1881. He received a B.S. degree in

JOEL P. WALLENSTEIN

electrical engineering from Pennsylvania State College in 1903.

He joined the Western Electric Company in 1903 and, in 1920, was assigned to the International Western Electric Company. He became a member of the International Telephone and Telegraph Corporation engineering department when, in 1925, it purchased the International Western Electric Company. Since that time he has continued in engineering work.

• • •

GEORGES PHELIZON was born in Paris on October 22, 1920. He graduated in 1938 from Ecole d'Electricité et Mecanique Industrielles de Paris, and from Ecole Supérieure d'Electricité in 1941.

He joined Laboratoire Central de Télécommunications in 1942 and has worked since then on circuit design in the microwave research department.

• • •

JOEL P. WALLENSTEIN was born on October 11, 1922, in Elizabeth, New Jersey. He received the B.S. degree in electrical engineering from the Newark College of Engineering in 1942. He then joined the Federal Telephone and Radio Corporation, Newark, New Jersey as a vacuum-tube engineer engaged in the development of high-power tubes. Since September, 1946, he has been engaged in the engineering consulting field.

Mr. Wallenstein is an Associate of the Institute of Radio Engineers and American Institute of Electrical Engineers, and a member of Tau Beta Pi.

• • •

For a biography and photograph of V. Belevitch, see Volume 24, Number 1, page 125.

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ELECTRICAL COMMUNICATION

Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies

ANGOLA RADIO NETWORK
PULSE-COUNT MODULATION
PULSE-TIME-MODULATION LINK FOR ARMY FIELD TELEPHONE SYSTEM
PROGRESS OF TELECOMMUNICATION SERVICES IN BRITISH POST OFFICE
SOJ-12 OPEN-WIRE CARRIER TELEPHONE SYSTEMS IN SOUTH AFRICA
OPERATIONAL CONTROL OF ELECTRICITY SUPPLY SYSTEMS
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DIMENSIONAL ANALYSIS APPLIED TO VERY-HIGH-FREQUENCY TRIODES
FLASHING SIGNAL FOR RAILWAY CROSSINGS
IN MEMORIAM—HENRY MARK PEASE

SEPTEMBER, 1947

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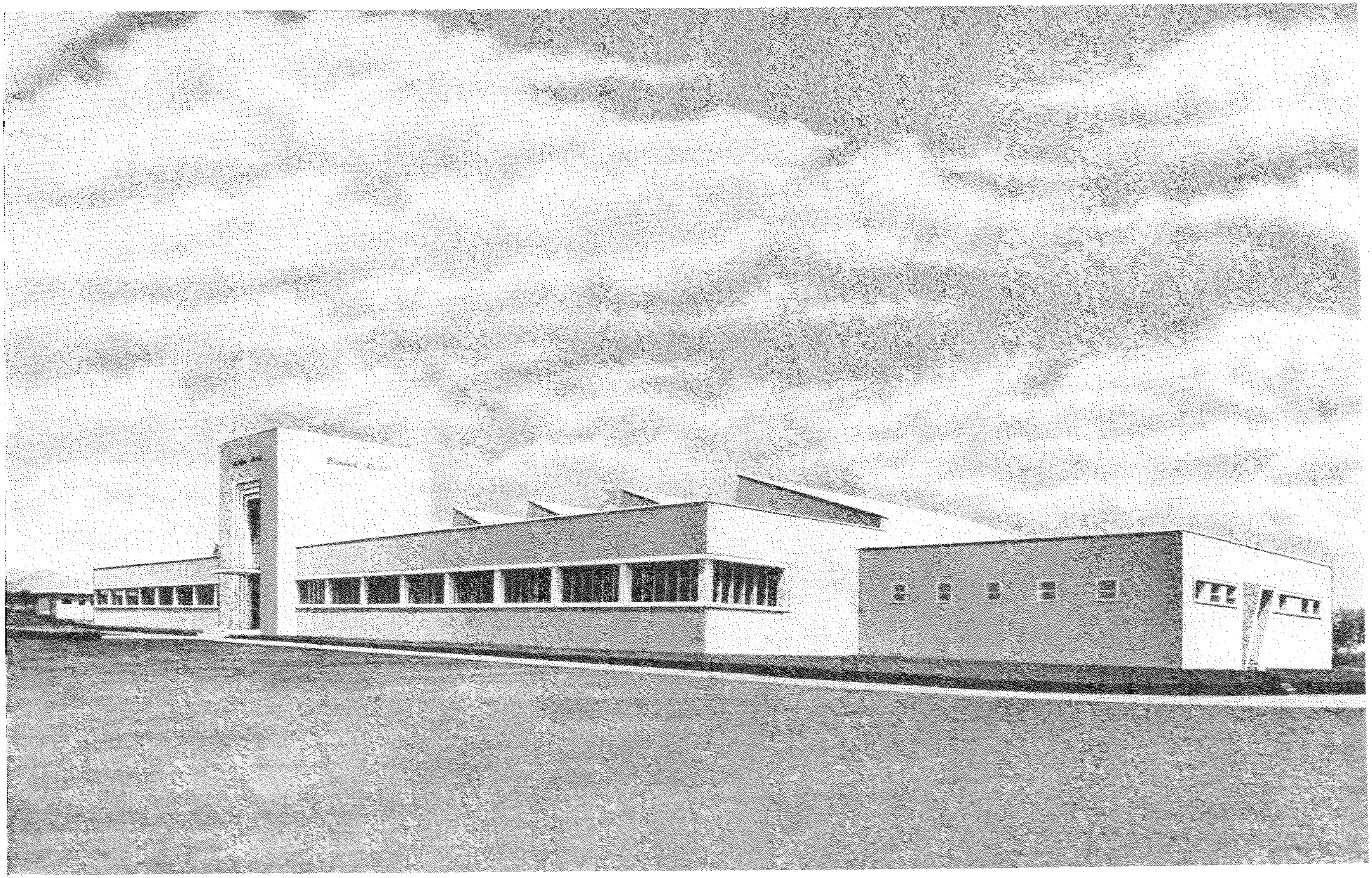


Compañía Standard Electric S.A.C. completed this new factory in Santiago, Chile, in December, 1945. In addition to the manufacture of radio and telephone equipment, the company also serves as installer for the Compañía de Teléfonos de Chile.

Angola Radio Network

By CARLOS PELAEZ

Standard Electrica, Lisbon, Portugal

ANGOLA, a Portuguese colony on the west coast of Africa, is situated between the Congo and the Cunene Rivers. It thus lies between the 6th and 18th southern degrees of latitude and is also between the 12th and 24th degrees of east longitude. The coastal province of Cabinda is an enclave in Belgian territory just a short distance north of the Congo River.

The coastal belt of Angola has a tropical climate. The interior regions, however, are generally at an altitude of between 2000 and 5000 feet above sea level and enjoy a more moderate climate. The highest region is in the Huila district in the southwest part where the city of Sá da Banderia is located. The total area of the colony is 1,250,000 square kilometers (480,000 square miles) with some 1000 kilometers (625 miles) of coastal belt. The white population numbers about 50,000, and there are approximately 3,700,000 natives.

Luanda is the capital and at the head of a railway running about 450 kilometers (280 miles) inland to Malange. The white population of Luanda is 8000. The second town of importance is Lobito, which is a terminal of the principal railway traversing the colony from west to east and joining Angola with the Belgian Congo and Elisabethville-Cape Town railway.

Angola exports cotton, sugar, palm oil, other agricultural products, and diamonds, the latter being an important factor in the economy of the colony.

1. Communications Within the Colony

The communications problem was to provide a service to link 21 specific places scattered throughout the colony as shown in Fig. 1. Economic and geographic conditions did not favor the use of physical circuits, and a radio system offered a practical solution of the problem. It was agreed that both radiotelegraph and radiotelephone service would be provided to the various centers, which are separated by distances varying between 140 and 700 kilometers (88 and 438 miles). The operation of the system has been adjusted to accord with the amount of traffic, and the power of the various transmitting installations reflects these conditions.

2. Communication With Ships

The coastal stations are designed to communicate with ships at sea as well as with the balance of the land network. The Luanda and Lobito stations are considered to have a range of 1000 nautical miles when operating in the band from 100 to 1000 kilocycles per second. In the same band, the Cabinda and Mossamedes stations are rated at 500 miles. The Santo Antonio do Zaire, Novo Redondo, and Baia dos Tigres stations carry but relatively small traffic and have a range of 300 miles.

All of these stations are, of course, equipped with high-frequency transmitters, which will permit much greater ranges in operating with ship stations and land stations in Angola.

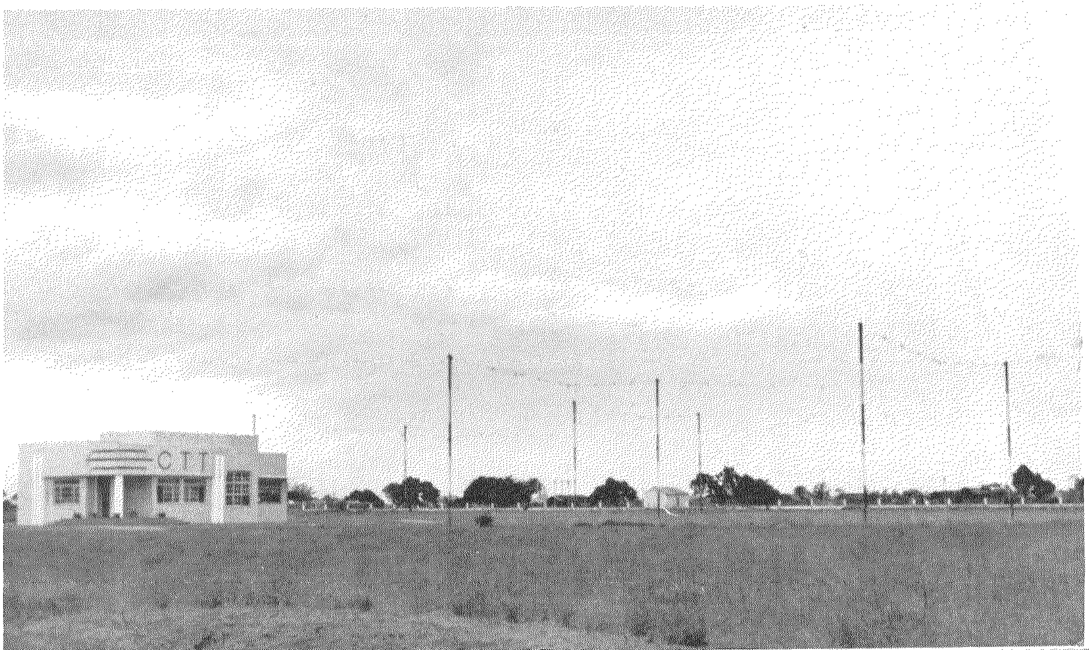
3. Aircraft Communication

In addition to the marine and overland communication facilities, all of the transmitters are arranged to permit operation at the frequencies assigned for aircraft communication.

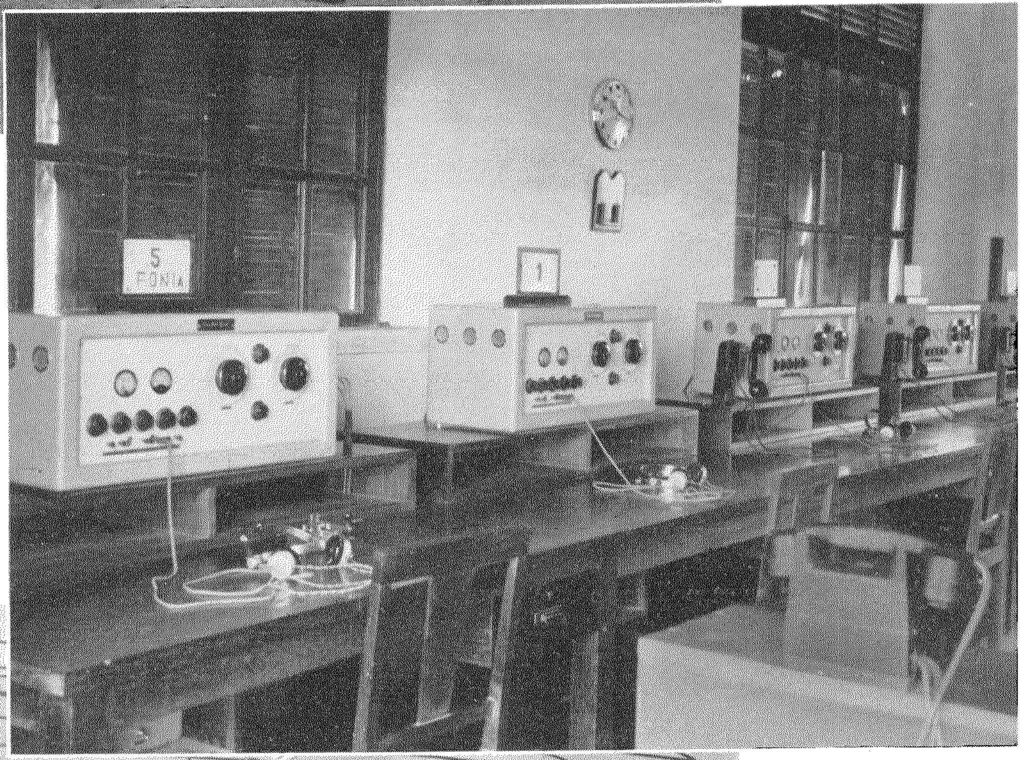
Luanda and Lobito are equipped with EL-4 transmitters having a rating of 1500/375 watts. Cabinda and Mossamedes use EL-1 transmitters rated at 500/125 watts. The transmitters at Santo Antonio do Zaire, Novo Redondo, and Baia dos Tigres are of the ESL-50 type of 500/125 watts. All high-frequency transmitters have been provided with a channel for aircraft communication.

4. Adcock Direction Finders

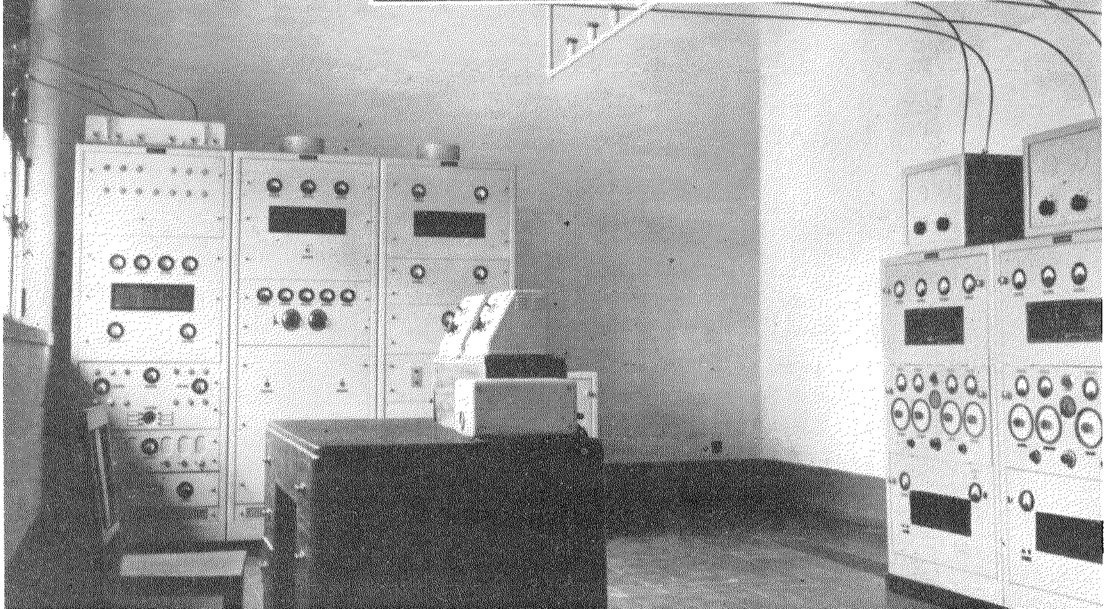
Six direction-finding installations of the Adcock type have been made near the airfields at Luanda, Cabinda, Lobito, Mossamedes,



Luanda, the capital of Angola, has higher-powered transmitters than any of the other stations. The receiving building is shown with the antenna systems used for both low and high frequencies.



The high-frequency receiving room at Luanda.



The transmitting room at Luanda showing the CS-2 transmitter in the background, the HS-1 equipment at the right, and the control desk in the foreground.

Nova Lisboa, and Vila Henrique de Carvalho. They are associated with the local radio stations. Further consideration is being given to the establishment of a special network for aeronautical communication.

5. *Equipment*

Despite wartime problems, the 21 stations have been completed. Each station includes the necessary transmitting and receiving equipment, power supplies, antennas, and essential terminating

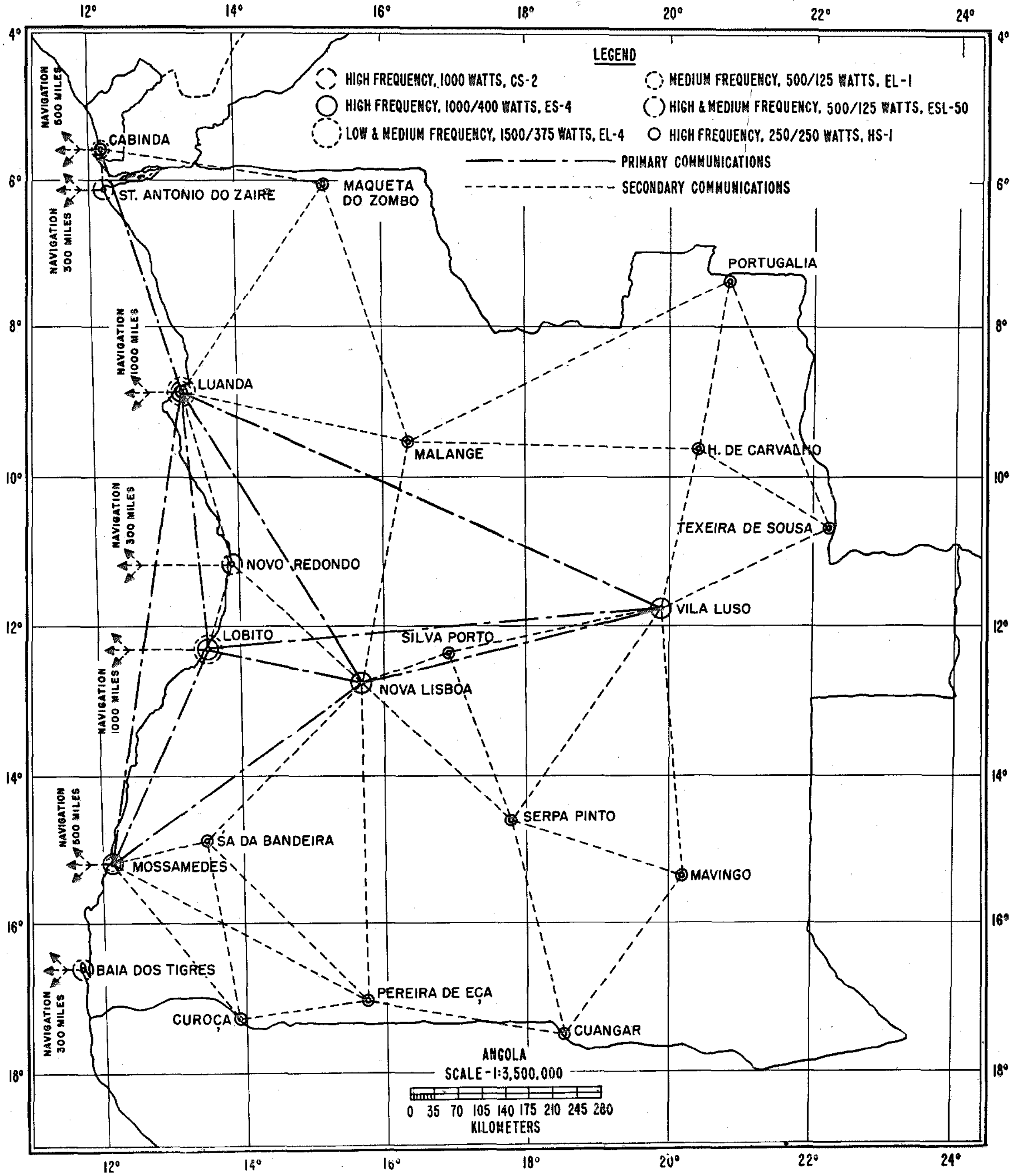


Fig. 1—Angola radio network linking 21 places scattered throughout the colony. Cabinda, on the northern coast, is separated from the main territory by part of the Belgian Congo and the Congo River. The coastal region is tropical and all equipment must be capable of operating in such a climate. The interior is at an elevation between 2000 and 5000 feet above sea level, which brings about a considerable improvement in the climate.

apparatus. All equipment was supplied by Standard Telephones and Cables, Limited, of London, with the exception of the diesel-engine-generator sets and power boards which were supplied by the British manufacturer, Ruston and Hornsby.

For the interior service, Luanda, Lobito, Nova Lisboa, Vila Luso, and Mossamedes are considered to be of greatest importance. At Luanda, a 1-kilowatt high-frequency transmitter type CS-2 is used to cover all points in the colony and for possible broadcasting services. Three HS-1 transmitters of 250-watt rating are provided for other traffic requirements. The other four towns are equipped with ES-4 transmitters rated at 1000/400 watts.

Santo Antonio do Zaire, Novo Redondo, and Baia dos Tigres are equipped with ESL-50 transmitters of 500/125 watts, which are adequate for their relatively minor amount of traffic.

All of the remaining stations have been equipped with 250/250-watt transmitters of the HS-1 type.

Terminal equipment of the TOP-9 type has been installed at Luanda to permit connection between the radiotelephone equipment and the local telephone network. At other places, this facility is provided by the use of 4-wire operation from a public telephone booth. Voice-operated carrier suppression quiets the transmitter during periods of reception. All of the principal stations have been equipped with B-2 privacy equipment.

6. Results

6.1 HIGH-FREQUENCY OPERATION

The specifications required that most stations be capable of communicating with the main station of each group. Tests have shown that any station in the network is capable of communicating with any other station, a condition that was specified only for the Luanda installation.

This results in much greater flexibility of operation and an increase in the traffic-carrying capacity as relaying from the smaller to the main stations is unnecessary.

Although nothing was included in the specifications regarding communication by high frequencies with places outside of the colony, Lobito maintained satisfactory contact with modulated continuous waves and telephony with Lourenço Marques at a distance of 2500 kilometers (1560 miles) and with planes in flight over any part of the colony as far as Lourenço Marques. Mossamedes is also able to communicate on modulated continuous waves and telephony with Lourenço Marques and with St. Thomas Island. Luanda, with a higher power than any other station, has communicated over even greater distances; it operates with Leopoldville, Pointe Noire, and St. Thame.

During the return voyage of the writer from Angola to Europe on the *S.S. Mousinho*, radiotelephone tests were made at distances up to 2000 kilometers (1250 miles) to Nova Lisboa and Lobito.

6.2 MEDIUM-FREQUENCY OPERATION WITH SHIPS

For operation with ships at sea, the specifications required a range of 1000 nautical miles for Luanda and Lobito, 500 miles for Cabinda and Mossamedes, and 300 miles for St. Antonio do Zaire, Novo Redondo, and Baia dos Tigres. Tests made from the first two stations with ships were successful over distances up to 2650 kilometers (1680 miles) and overland with Tanganyika at 2160 kilometers (1350 miles). Lobito and Mossamedes maintained communication with the *S.S. Mousinho* on a voyage from Lourenço Marques to Lobito.

Although no specifications were included on radiotelephone service for these stations, tests so far carried out from Mossamedes, which is required to have a range of 500 miles on radiotelegraphy, have resulted in communication well over 900 kilometers (560 miles) on radiotelephony.

7. Wartime Difficulties

The chief difficulties encountered resulted from the outbreak of war and affected not only the supply of equipment from England but also the means of transportation both to the colony and within Angola. Deficiencies in staff were similarly encountered.

Despite these handicaps, 21 complete stations were erected and include 25 transmitters, 51 receivers, steel antenna masts, cables, motor-generator sets (some in duplicate), together with terminal equipment, dry-type rectifiers, storage batteries, and a large quantity of auxiliary material. All of this equipment and supplies were provided by Standard Telephones and Cables, Limited.

Pulse-Count Modulation

By D. D. GRIEG

Federal Telecommunication Laboratories, Inc., New York, N. Y.

ADVANTAGES of pulse transmission include the use of time-division multiplexing, absence of cross talk introduced by nonlinear circuit elements, and improved transmission characteristics. Modulating methods may involve variation of such pulse characteristics as amplitude, frequency, width, and timing. Pulse-count modulation, sometimes called pulse-code modulation, is based on the printing-telegraph binary code to indicate the amplitude of the modulating wave from instant to instant. Thus, to the discrete sampling of a modulating wave in a time dimension is added a discrete sampling of the amplitude of the wave. Electronic methods of coding and decoding the instantaneous amplitudes of audio-frequency modulating waves are described. The system permits the advantages of the printing telegraph to be extended to voice transmission. Substantial improvements in signal-to-noise ratios, cross talk, threshold values, and other transmission factors are indicated. These improvements have been verified under operating conditions.

• • •

Within recent years, a wide variety of modulation methods utilizing short bursts or pulses of energy have been developed. These methods have found application to a large number of systems including telephony, radio relaying, telemetering, and broadcasting.[1–6] The particular properties that have made pulse modulation attractive for transmission systems have resulted from both its multiplexing and modulation characteristics.

For example, by interleaving the modulating signals in time sequence, i.e., time-division multiplex, cross talk introduced by nonlinearities in the transmission system, such as would be obtained with frequency-division multiplexing, is eliminated. This important property allows the use of circuit elements with nonlinear characteristics and makes possible long relays with many repeaters. A further factor obtained with constant-amplitude pulse systems is independence of fading and other transmission vagaries. An additional property is a flexibility that allows various transmission parameters to be exchanged, for example, bandwidth for noise-reduction properties, and distortion and cross talk for bandwidth. This permits systems to be designed to specific requirements.

Basically, pulse transmission involves sampling the modulating signal at discrete intervals of time sufficiently short so as to allow little or no change in the modulating signal during the period of sampling. As these samples or pulses may be characterized by the parameters of timing, duration, frequency, build-up time, decay time, and shape, a large number of modulation methods involving these quantities either singly or in combination may be envisaged.

The amplitude of the pulses may be varied resulting in pulse-amplitude modulation (PAM). Alternatively, the width or duration of the pulses can be made to vary with the modulating signal resulting in pulse-width modulation (PWM). Instead of varying the individual pulse characteristics, the time between pulses, or with reference to a marker pulse, can be varied, resulting in pulse-time modulation (PTM). Or the repetition frequency of the pulse can be varied, comparable to conventional frequency modulation, yielding pulse-frequency modulation (PFM). Other characteristics of the pulse, such as the build-up or decay time, can be varied. Combinations of these several modulations may be employed to produce various hybrid systems.

---

[1] E. M. Deloraine and E. Labin, "Pulse Time Modulation," *Electrical Communication*, v. 22, n. 2, pp. 91–98; 1944.

[2] F. F. Roberts and J. C. Simmonds, "Multichannel Communication Systems," *Wireless Engineer;* November, 1945.

[3] D. D. Grieg and A. M. Levine, "Pulse-Time-Modulated Multiplex Radio Relay System—Terminal Equipment," *Electrical Communication*, v. 23, pp. 159–178; June, 1946.

[4] H. S. Black, J. W. Beyer, T. J. Grieser, and F. A. Polkinghorn, "A Multichannel Microwave Radio Relay System," *Electrical Engineering*, v. 65, pp. 798–805; December, 1946.

[5] L. L. Rauch, "Electronic Commutation for Telemetering," *Electronics;* February, 1947.

[6] D. D. Grieg, "Multiplex Broadcasting," *Electrical Communication*, v. 23, pp. 19–26; March, 1946.

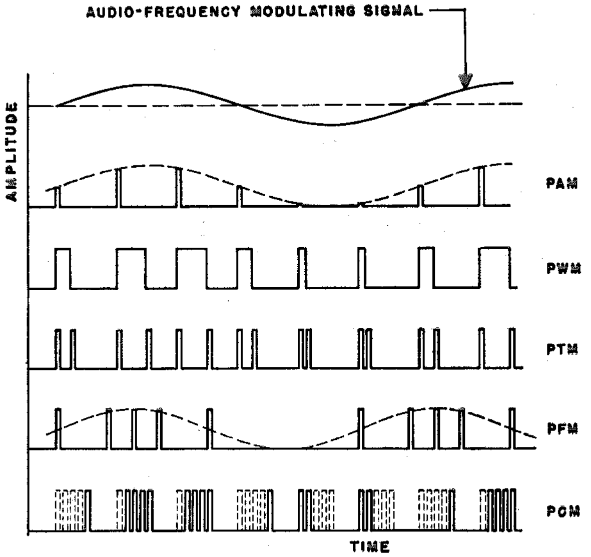


Fig. 1—Pulse-amplitude, pulse-width, pulse-time, pulse-frequency, and pulse-count modulations are illustrated.

Fig. 1 illustrates the general characteristics of a few of the possible modulation methods. It should be noted that the figure illustrates only the so-called "video-frequency" pulses, which can be used directly for transmission over wires or cables or to modulate the radio-frequency carrier in either amplitude, frequency, or phase.

1. Pulse-Count Modulation

With most methods of pulse modulation, all levels of the modulating signal between zero and maximum are transmitted in the sampling process. However, a second class of modulation can be devised in which, in addition to time sampling, only selected levels of the amplitude are transmitted; i.e., a system of double discreteness involving quantization of amplitude at discrete intervals of time. With such a system, the amplitude range of the modulating signal is divided into a number of discrete levels. If the instantaneous amplitude of the signal falls between two levels, either the lower or upper level is transmitted depending on which is closer to the signal amplitude. These methods were first described by A. H. Reeves in United States and French patents.[7]

Because only a finite number of levels is involved, it is possible to transmit the modulating information by a code similar to that of the printing telegraph system. For example, if the modulating signal is divided into a total of 31 levels, a five-unit binary numbering system may be used to identify each discrete amplitude. Thus, all levels from 0 to 31 would be transmitted in terms of 0 and unity, which for practical purposes may be the absence and presence of a pulse or any other two-value variation such as a difference in frequency. 0 is transmitted as 00000, 1 as 10000, 2 as 01000, 3 as 11000, and 31 as 11111. Fig. 1 compares this system of transmission with other methods. Fig. 2 gives the binary count combinations utilizing five pulses to reproduce any level between 0 and 31.

	WEIGHT OF PULSE				
LEVEL NUMBER	1	2	4	8	16
1	X				
2		X			
3	X	X			
4			X		
5	X		X		
6		X	X		
7	X	X	X		
8				X	
9	X			X	
10		X		X	
11	X	X		X	
12			X	X	
13	X		X	X	
14		X	X	X	
15	X	X	X	X	
16					X
17	X				X
18		X			X
19	X	X			X
20			X		X
21	X		X		X
22		X	X		X
23	X	X	X		X
24				X	X
25	X			X	X
26		X		X	X
27	X	X		X	X
28			X	X	X
29	X		X	X	X
30		X	X	X	X
31	X	X	X	X	X

Fig. 2—Binary count combinations. By adding the weights of the pulses for which there are *X*'s, the level number will be obtained. In transmission, the *X*'s correspond to the presence of pulses.

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[7] A. H. Reeves, U. S. Patent 2,272,070, February 3, 1942; also French Patent 852,183, October 23, 1939.

In Fig. 3, the solid line represents the audio-frequency modulating voltage. This signal is broken into discrete levels resulting in the step-type function also shown. At the time of scanning, the instantaneous level of the step function is reproduced in terms of the five-unit pulse code likewise shown. The type of modulation combining both time sampling as well as discrete amplitude or quantization has been termed pulse-count (also pulse-code) modulation.[8]

This system permits many advantages of the printing telegraph to be extended to voice transmission.

For example:

A. Noise added to the system can be made noncumulative for a long relay using many repeaters because each pulse can be completely regenerated at each repeater with complete suppression of noise previously added in the transmission path, provided the minimum signal received is above the noise threshold. This allows a considerably larger amount of attenuation over other modulation methods, which in practice permits greater distance between repeaters, smaller power requirements, and greater freedom from fading variation.

B. Where multiplexing of the pulse-count signal is by means of time division, cross talk is noncumulative. This factor is of considerable importance because the effect of reflections resulting from multipath transmission in a radio system or, alternatively, caused by mismatches or discontinuities in cable transmission, can be minimized.

C. The system operates effectively with a signal approximately 15 decibels above thermal-agitation noise or 3 decibels above peak noise, in addition to an extremely large signal-to-noise improvement ratio, which is substantially independent of the signal-to-noise input ratio when that ratio exceeds 15 decibels.

D. The transmission method also allows relatively simple repeaters to be used because only on-off characteristics need be recognized and transmitted.

2. Technical Properties

2.1 Quantization Distortion

A consideration of first importance is the distortion[9] introduced by quantization of amplitude and sampling in time. It is obvious that, as the number of levels is increased, the granulation of the signal becomes smaller and hence distortion

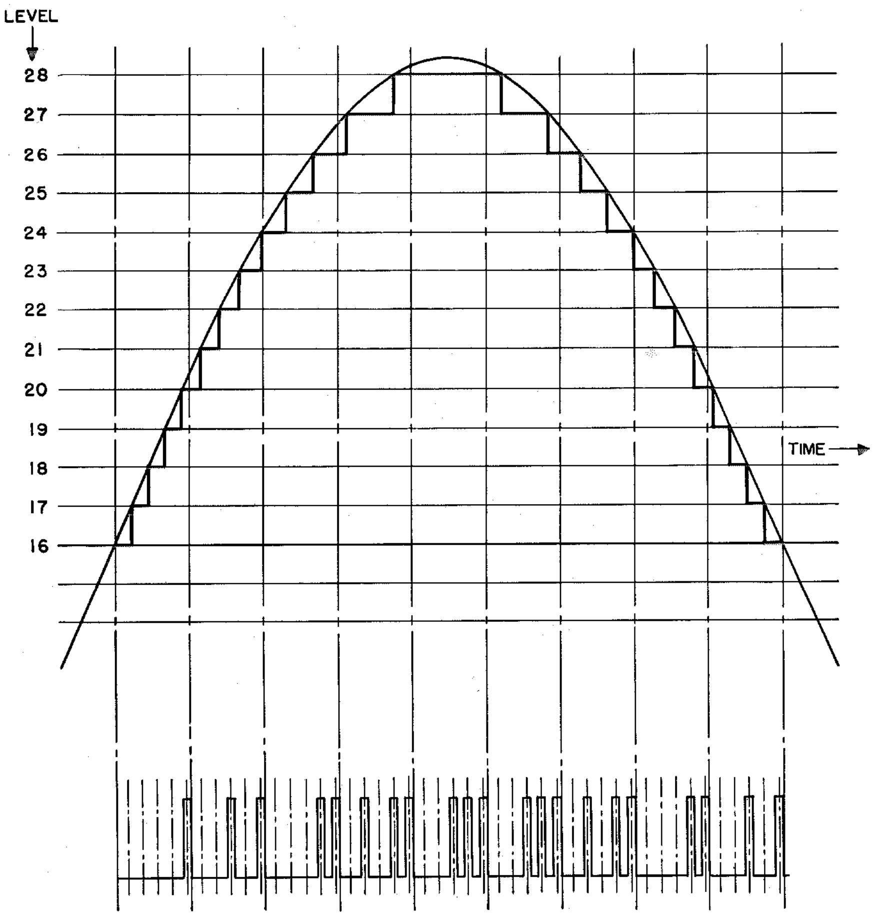

Fig. 3—Details of pulse-count modulation. The solid continuous line represents the audio-frequency modulating voltage, and the step function corresponds to the quantized signal levels. The pulse-count signal at the bottom of each column corresponds to the value of the level at the left side of that column.

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[8] H. S. Black and J. O. Edson, "Pulse Code Modulation," American Institute of Electrical Engineers Summer Convention, Montreal, Quebec, Technical Paper 47–131; June, 1947.

[9] A. G. Clavier, P. F. Panter, and D. D. Grieg, "Distortion in a Pulse-Count Modulation System," American Institute of Electrical Engineers Summer Convention, Montreal, Quebec, Technical Paper 47–152; June, 1947.

resulting from this factor is decreased. On the other hand, this requires an increase in the number of code pulses to be transmitted and a corresponding increase in bandwidth. The reference cited shows that the use of a relatively small number of levels is satisfactory for generally acceptable speech quality. For example, dividing the speech waveform into a total of 31 levels yields a distortion of the order of 3 percent. Tests have been conducted which indicate that intelligible speech can be obtained with as low a number as 7 levels.

The percent distortion D is a function of the total number of levels $2m$, where m is the number of levels on one side of the zero axis, and is given by

$$D=\frac{1}{\sqrt{6}(m)}\times 100. \qquad (1)$$

This expression holds reasonably well for a number of levels less than aproximately 100. Table I gives the distortion for various numbers of levels into which the signal may be divided.

2.2 Time-Sampling Distortion

Time-sampling distortion[10] results from scanning the signal at discrete intervals of time. Time-sampling distortion for pulse-count modulation is the same as for pulse-amplitude modulation; i.e., spurious components are developed mainly as a result of sideband harmonics of the pulse carrier falling into the audio-frequency transmission band. There is, in addition, a small distortion component caused by the finite width of the sampling pulse. For normal pulse widths and for practical purposes, this source of distortion can be neglected.

The minimum ratio of sampling frequency f_p to audio-frequency bandwidth f_h-f_l, where f_h and f_l are the high- and low-frequency limits of the audio-frequency band, respectively, is given by

$$\frac{f_p}{(f_h-f_l)}=2. \qquad (2)$$

It should be noted that in this case, in order to achieve the maximum sampling frequency, it may be necessary to transpose the band (f_h-f_l) to a band having a top frequency of f_h.

TABLE I

Distortion as a Function of Number of Levels

Number of Levels	Distortion in Percent
3	27
7	13
15	7
31	3.5
63	1.7
127	0.8
255	0.4
511	0.2
1023	0.1

TABLE II

Binary Count System

Number of Levels as a Function of Number of Pulses

Number of Pulses	Number of Levels
1	2
2	3
3	7
4	15
5	31
6	63
7	127
8	255
9	511
10	1023

For practical purposes, a larger ratio must be utilized to permit the sampling components to be separated from the voice components with an economical audio-frequency filter. With a simple filter, a cutoff frequency $1.5f_l$ can be attained. Also, if the lower cutoff frequency is small compared to the upper frequency, (2) becomes

$$\frac{f_p}{f_h}\cong 2.5. \qquad (3)$$

2.3 Number of Pulses

Preferably, the quantized signal is transmitted by a series of pulses as previously mentioned. If p is the total number of pulses to be transmitted for any given level, the number of levels $2m$ is given by $2m=N^p$, where N corresponds to the counting system used. If a binary count system corresponding to an on-off function is utilized, $N=2$ and $2m=2^p$, or

$$p=1.44\log_e 2m. \qquad (4)$$

Note that the number of pulses must be the nearest whole number to the value given by (4). Table II illustrates the number of pulses required for the various numbers of levels utilizing a binary count system.

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[10] H. L. Krauss and P. F. Ordung, "Distortion and Bandwidth Characteristics of Pulse Modulation," American Institute of Electrical Engineers Summer Convention, Montreal, Quebec, Technical Paper 47–166; June, 1947.

2.4 Bandwidth Considerations

The transmission of pulses obviously requires a larger bandwidth than that normally employed for voice transmission. On the other hand, with pulse-count modulation, it is necessary to determine only the presence or absence of a pulse and, hence, a comparatively smaller bandwidth than that required for other types of pulse systems is satisfactory. Several factors determine the allowable pulse distortion and the corresponding bandwidth. For example, the amount of carry-over from one pulse to the adjacent pulse determines the cross talk in the system. This same factor influences the signal-to-noise ratio and noise threshold. Alternatively, the bandwidth must be properly defined because the pulse distortion is a function of the low and high cutoff frequencies, and the type and rate of change of attenuation at these cutoff points, as well as of the bandwidth itself.

The minimum bandwidth required for zero cross talk has been determined both theoretically and empirically for various types of cutoff characteristics including the frequency characteristics of equalized and unequalized coaxial cable. The results can be expressed approximately as follows:

$$F \cong \frac{f_p}{2} \tag{5}$$

for the upper limit of frequency, where F is defined as the pulse bandwidth at the 3-decibel points and f_p is the total number of pulses per second.

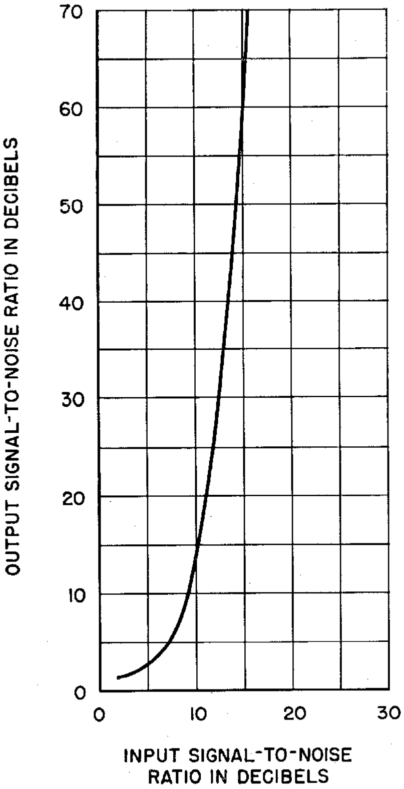

Fig. 4—Output signal-to-noise ratio plotted against input signal-to-noise ratio for pulse-count modulation.

2.5 Signal-to-Noise and Threshold

Pulse-count modulation, in common with the printing telegraph system, yields a far greater signal-to-noise improvement than any other modulation system using equal bandwidth. With conventional amplitude, frequency, or pulse-time modulation, audio-frequency output signal-to-noise ratio is always directly proportional to the input carrier-to-noise ratio, provided the carrier is above the threshold value. With pulse-count modulation, however, the output signal-to-noise ratio is essentially independent of the input ratio once the signal exceeds the noise level.

Noise enters into a pulse-count system only if a noise pulse substitutes for, or suppresses, a pulse of the transmitted series. Under conditions where noise never exceeds approximately half the peak pulse amplitude, it is always possible to "slice out" a portion of the signal pulse that is completely undisturbed by noise. The time of occurrence of this pulse may be advanced or retarded but, unlike other pulse systems, this effect is unimportant because it is necessary to determine only the presence or absence of a pulse. Thus, it would be assumed that for signals above twice the noise level, the output signal-to-noise ratio is essentially infinite.

Noise has a random distribution of peak amplitudes. This varying threshold requires that the input signal must pass through a range of values before a maximum output signal-to-noise ratio is achieved. Various calculations and tests indicate that a root-mean-square value of signal-to-noise of the order of 15 decibels gives useful service. This is illustrated by the graph of Fig. 4. This threshold value of 15 decibels has also been observed experimentally in teleprinter tests to correspond to the "breaking point" between perfect and imperfect reception.

2.6 General Design Relations

The expressions for the number of pulses in relation to the number of levels (4) and for distortion as a function of the number of levels (1) can be combined to give the relation between the number of pulses and distortion.

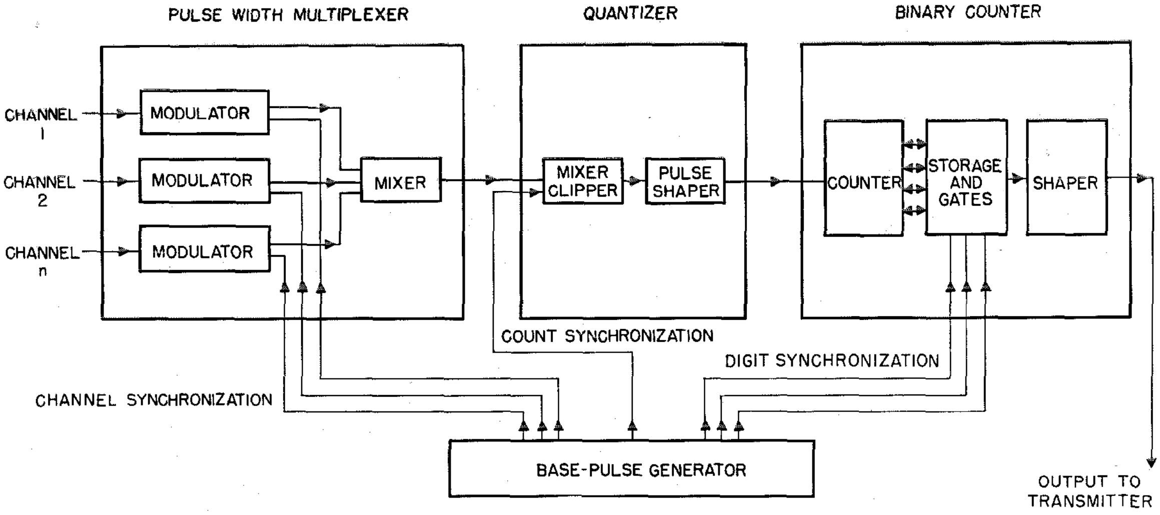

Fig. 5—Block diagram of pulse-count modulator. This unit transposes the modulation of the individual channels to a multiplex series of pulses.

$$p = 1.44 \log_e \frac{80}{D}. \tag{6}$$

The ratio of pulse frequency to audio frequency is given for the practical case by (3) and the number of pulses per second per channel is given by

$$f_p = 3.6 f_h \log_e \frac{80}{D}. \tag{7}$$

For N channels, this expression is, of course, multiplied by N. The frequency band is given in (5) as the total number of pulses divided by 2 and, therefore, the ratio of the required transmission bandwidth to the audio-frequency bandwidth is

$$\frac{F}{f_h} = 1.8N \log_e \frac{80}{D}. \tag{8}$$

For a distortion D of 3 percent, which is a reasonable practical value for telephony and corresponds to a 31-level system, we obtain

$$\frac{F}{f_h} = 6.25N. \tag{9}$$

It should be noted that this corresponds to the case for a simple filter for separating the audio-frequency components from the pulse scanning components. Theoretically, if the ratio of pulse frequency to audio-frequency bandwidth $f_p/f_h = 2$ is utilized, the above expression becomes

$$\frac{F}{f_h} = 5N. \tag{10}$$

In other words, the theoretical bandwidth for a 31-level pulse-count system is five times that of a single-sideband amplitude-modulation system.

3. *Equipment*

Several methods are available for producing pulse-count modulation and for its demodulation. A representative method has been described by A. H. Reeves[7] and the modulating system is illustrated by the block diagram of Fig. 5. This corresponds to a multichannel pulse-count-modulation system. Although applicable to a large number of channels, for simplicity only a small number is illustrated.

3.1 Modulator

The modulator serves both to multiplex the individual audio-frequency channels as well as to translate the channel information into a series of coded pulses. It consists of four main units: (A) pulse-width multiplexer, (B) quantizer, (C) binary counter and associated circuits, and (D) base-pulse generator.

The modulating voltages applied to the channel modulators operate on the timing pulses supplied by the base-pulse generator to yield width-modulated pulses occurring in proper time sequence. They yield a pulse-width multiplex series at the mixer output as illustrated in Fig. 6.

This series of pulses passes to the quantizer circuit. Each individual pulse allows the passage of a number of count pulses corresponding to the width of that pulse. The maximum number of count pulses corresponds to the maximum number of levels and is produced only by a pulse of maximum width.

The series of pulses obtained from the quantizer must then be translated into a binary number in the binary counter shown. The counter can take many forms, the most familar being that of a series of "flip-flop" multivibrators interconnected so that each multivibrator turns over once for every two pulses applied by the preceding multivibrator. Associated with the binary counter are the necessary storage and gate circuits in addition to reset circuits.

A series of digit synchronizing pulses, which have the same time characteristics as the ultimate transmitted pulses, are applied to the gates in addition to the final voltage yielded by each individual multivibrator at the end of the count. Digit pulses are passed by the gates only when a potential corresponding to a full turnover of the individual multivibrator exists.

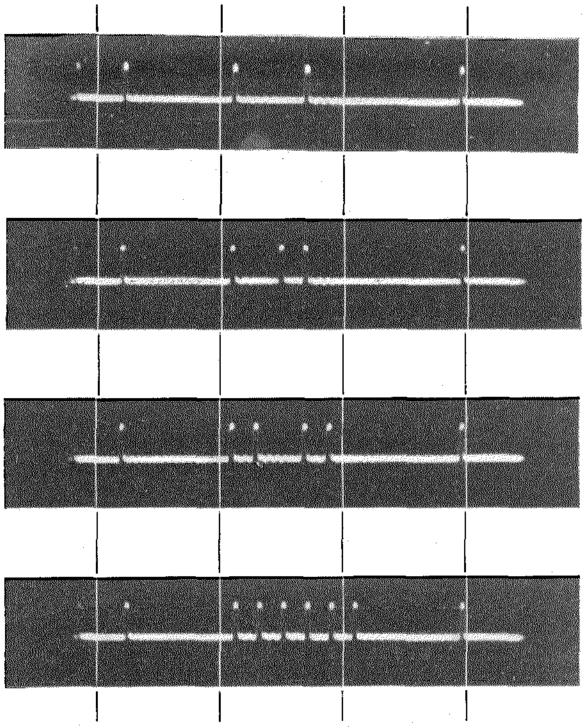

Fig. 7—Oscillogram of pulse-count-modulation signals. The counts set up for the center channel are 9, 13, 27, and 31 respectively.

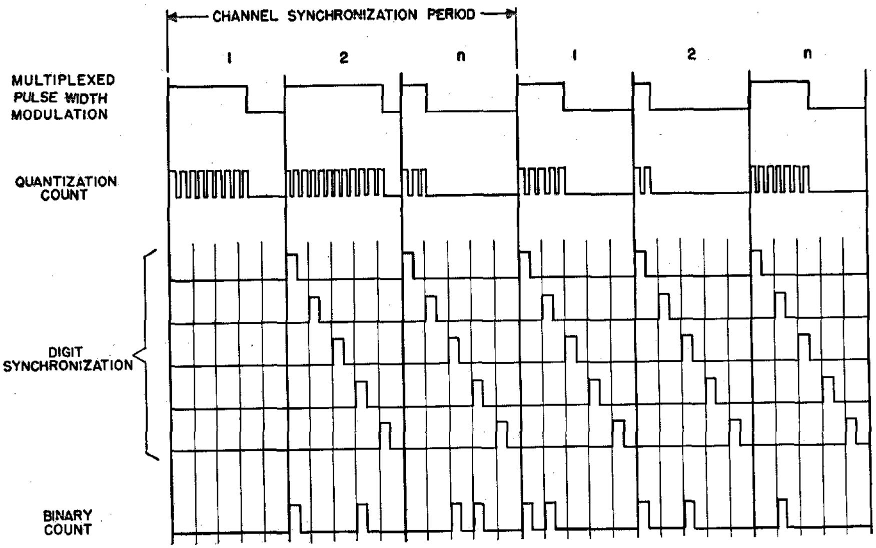
CHANNEL SYNCHRONIZATION PERIOD
1
2
n
1
2
n
MULTIPLEXED PULSE WIDTH MODULATION
QUANTIZATION COUNT
DIGIT SYNCHRONIZATION
BINARY COUNT

Fig. 6—Modulator waveform sequence. The binary count is delayed as it must follow completion of the quantization count. The binary count thus appears in the next succeeding channel.

The storage circuit is necessary to provide a full count of the instantaneous level prior to setting up the ultimate binary count. The reset signals are, of course, utilized to recycle the counters for subsequent counting. As an example of the pulse rates utilized, the data of Table III are representative for an 8-channel, 31-level pulse-count-modulation system with channel bandwidths of 3500 cycles.

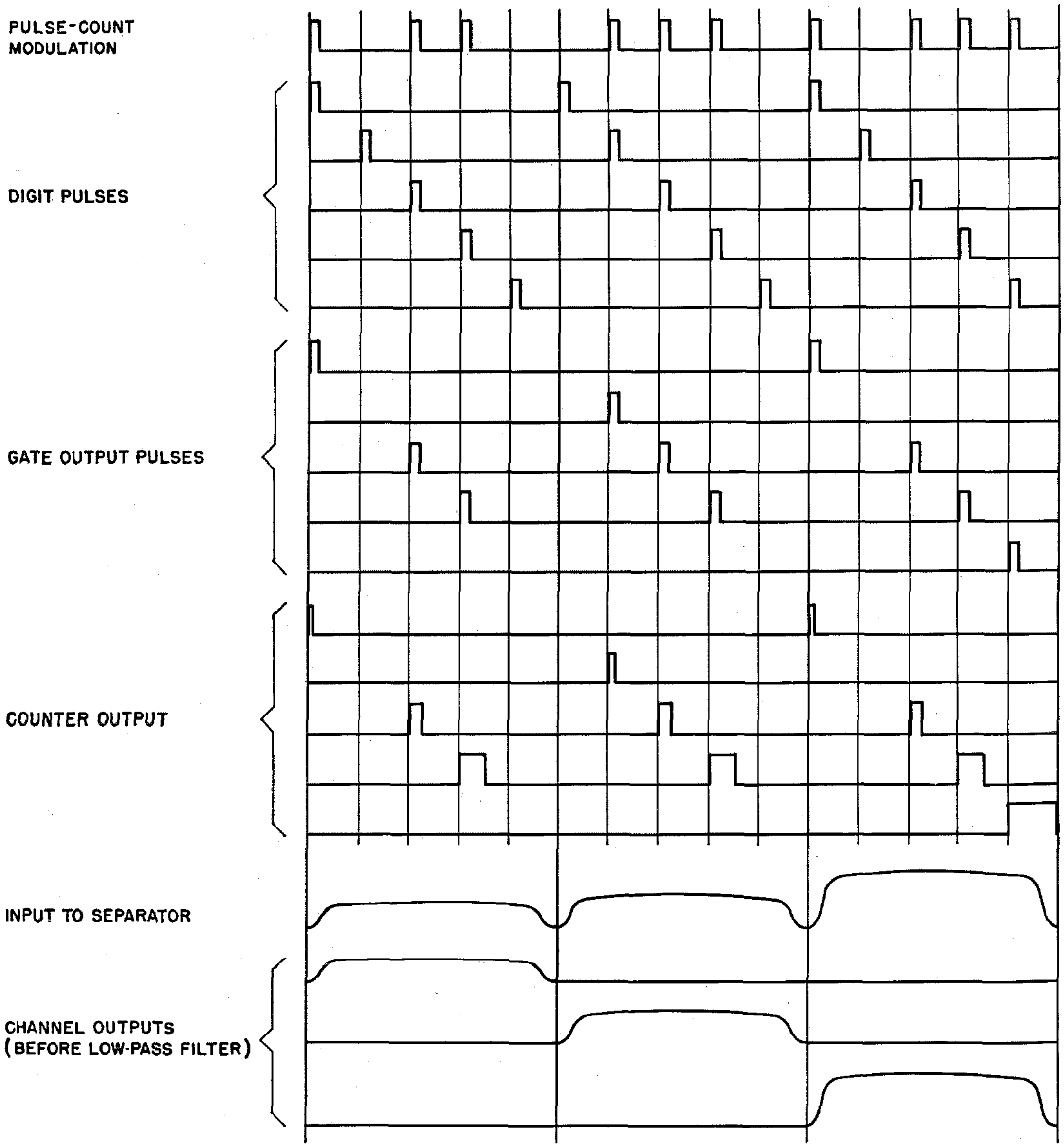

Fig. 9—Demodulation waveform diagram. Audio-frequency filters, not shown in the block diagram, are utilized to remove the pulse-frequency components from the modulating-signal output of the separator-demodulator units.

The final output pulses corresponding to proper binary count for each sample are then obtained from the appropriate shaper circuit for transmission. An oscillogram illustrating the pulses derived from a system similar to that described is shown in Fig. 7.

TABLE III

Representative Pulse Rates

8 Channels, 31 Levels

	Kilocycles
Base Pulse Rate (Channel Synchronization)	8
Count Synchronization	1000
Digit Synchronization	333
Output Pulse-Repetition Frequency	333

3.2 Demodulator

The demodulator accepts the multiplex pulse series, separates the individual channels, and recovers the original modulation. The system illustrated in Fig. 8 is essentially the reverse of that described for the modulator. As indicated by the illustration, the receiver consists of three main units: (A) synchronizing circuit, (B) counter circuit, and (C) multiplex demodulator.

The synchronizing circuit extracts from the incoming pulses information for producing a pulse series for the various controls, such as for digit synchronization, channel order, and channel synchronization. The input signal is obtained from a radio receiver and is in the form of a series of pulses. This signal is applied to gate circuits which, in conjunction with the digit synchronizing voltages, separate the code series into individual pulses corresponding to the actual digits transmitted. This is illustrated in the waveform diagram of Fig. 9. If a code pulse and digit pulse occur simultaneously, the gate circuits produce output. The individual digit pulses actuate a counter circuit that produces a pulse whose width corresponds to the weight of the digit applied. Thus for a 5-pulse system, a total of 5 counters would be used with each counter producing successively an output pulse twice the width of the preceding counter. In this example, the pulse width produced by the fifth counter would be sixteen times the width of the first counter; preceding counters would produce pulse widths having the relative weights of 8, 4, and 2, respectively.

The outputs of the counters are connected in parallel to obtain the sum of the counter pulses produced. This signal is then passed through a filter which removes the high-frequency components. It is then applied to the multiplex demodulator, which serves to separate the individual channels and translate the energy variation into the appropriate audio-frequency signal.

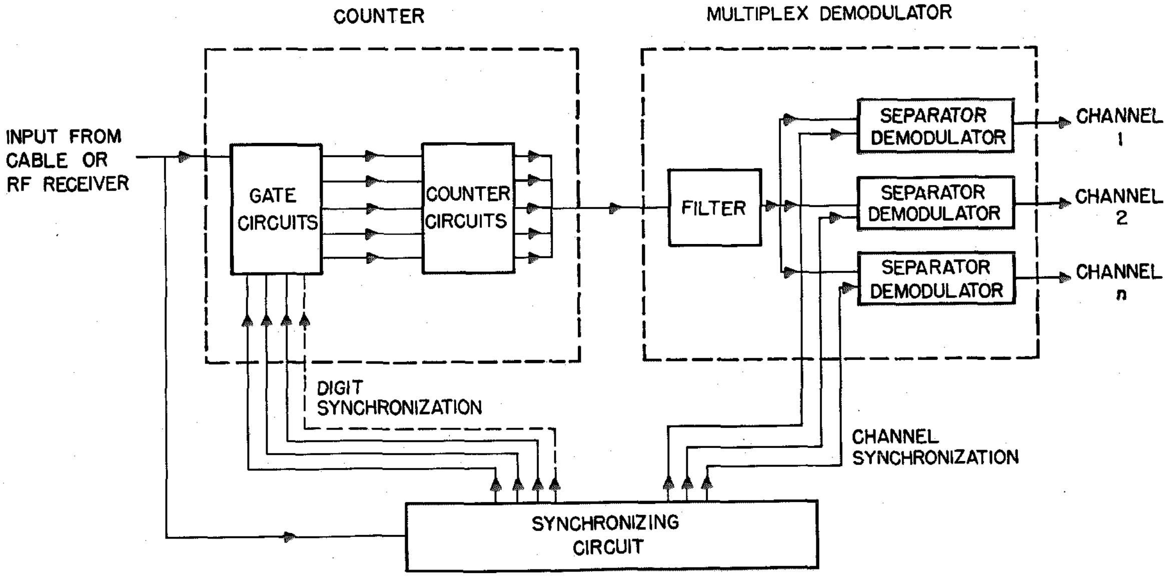

Fig. 8—Block diagram of receiving demodulator. This unit separates the pulse-count multiplex signal into individual channels. In addition, the pulse-count combinations are translated into the corresponding audio-frequency signals.

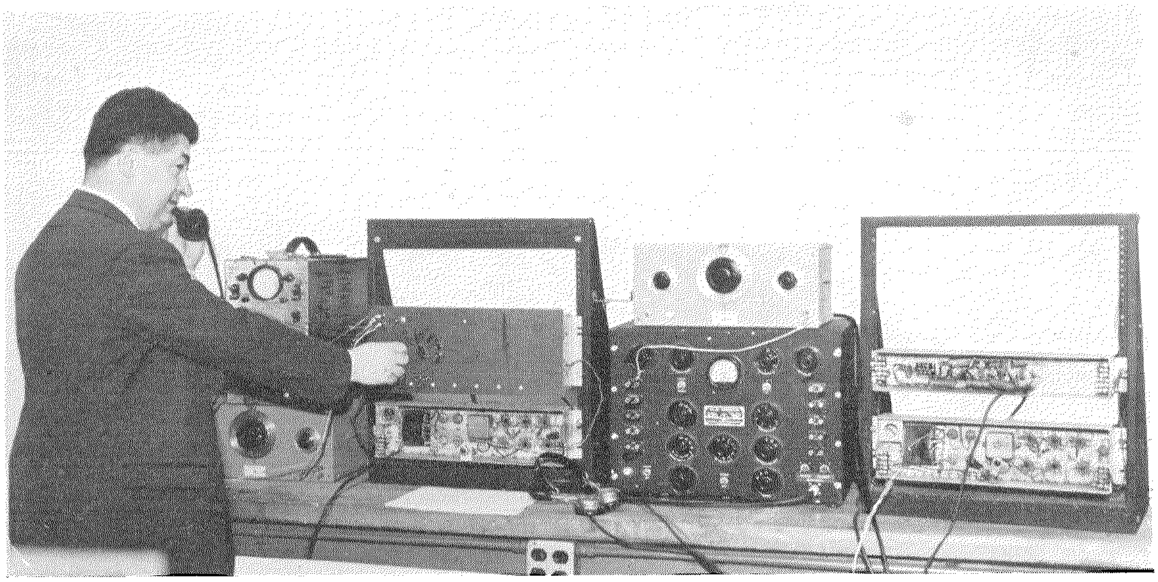

Fig. 10—Representative experimental system for pulse-count-modulation transmission. Included in this apparatus are both the modulator and demodulator for transmission and reception.

Simplified systems of pulse-count-modulation transmission corresponding in general to that described have been constructed in the laboratory. Fig. 10 illustrates such a simplified system developed for experimental purposes.

4. Conclusion

Tests of pulse-count-modulation transmission have tended to confirm the advantages theoretically indicated. In particular, the telegraphy-type characteristics of the system have permitted operation over relatively unfavorable transmission paths, such as poor cable, without destruction of the signal-carrying characteristics. A similar attractive characteristic can be expected over radio transmission paths where severe fades, as well as multipath reflection, are to be expected.

On the basis of the experimental results obtained to date, pulse-count modulation would seem to offer attractive possibilities for application to radio and wire transmission circuits, particularly for multichannel operation over long relay paths. These applications are being investigated.

Pulse-Time-Modulation Link for Army Field Telephone System

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DURING the war, a need developed for an apparatus that could be introduced between two field telephones to permit them to function normally without a pair of wires connecting them. This type of operation can be achieved by using a duplex radio link and suitable arrangements for changing from the 2-wire telephone circuit to the 4-wire system needed for the radio link.

When pulse modulation is used, the pulses transmitted by each station may be radiated at times when no incoming pulses are due. This form of time division permits duplex operation.

Advantages of this method of duplexing include the use of a single channel for both directions of transmission, increased difficulty of unauthorized interception, and the use of a common antenna system for transmission and reception. An increased rejection of noise in the radio link results from the use of high peak powers which characterizes pulse methods.

1. Principles

The equipment used at one terminal of such a link is shown in block form in Fig. 1. Most of the elements of the system are identical with those used for conventional pulse-time-modulation voice communication without duplexing. The only additions for duplexing are the sine-wave generator and phasing control connected to the output of the receiver and the hybrid network needed to convert from 2-wire to 4-wire operation.

In operation, either station may be used as the "master" station to establish the pulse-repetition rate for both terminals of the link. The pulse-repetition rate is controlled by a sine wave generated by a stable oscillator in the sine-wave-generator and phaser chassis. This wave is shaped by a pulse-time modulator[1] into a train of pulses, with small deviations from the normal positioning representing the voice modulation. These pulses are amplified and applied as modulation to the transmitter. The modulator unit also forms pulses to block the intermediate-frequency amplifier of the receiver during the time a pulse is to be

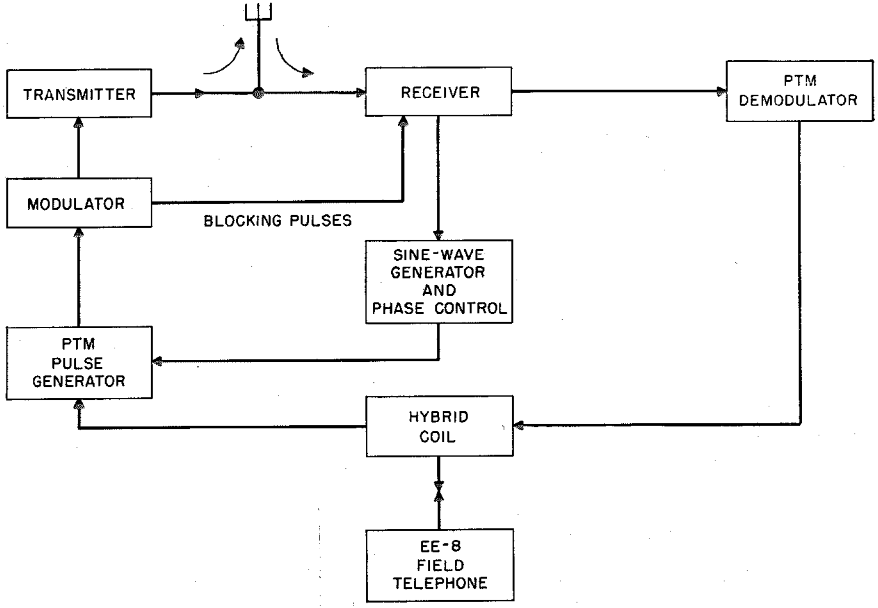

Fig. 1—Block diagram of pulse-time-modulation link used to extend a field telephone system.

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[1] D. D. Grieg and A. M. Levine, "Pulse-Time-Modulated Multiplex Radio Relay System—Terminal Equipment," *Electrical Communication,* v. 23, pp. 159–178; June, 1946.

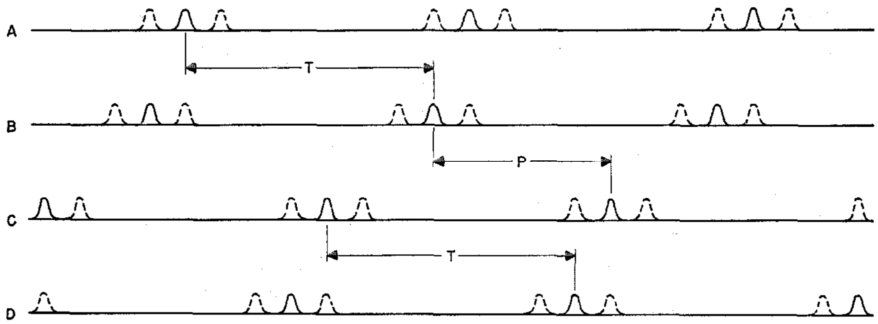

Fig. 2—Relative timing of pulses. *A*. Pulses as emitted from the master station. The dotted pulses indicate the extreme positions assumed under modulation. *B*. Reception of pulses at the slave station. *T* is the time required for transit from the master station. *C*. The pulses are emitted by the slave station after further delay *P* introduced by the phase control at the slave station. *D*. Pulses from the slave station are received at the master station suitably interleaved with those transmitted by the master station, thus allowing duplex operation without interference.

transmitted, thus preventing overload and cross talk in the receiver. The received pulses are demodulated in the usual way and pass through the hybrid circuit to the field telephone line.

At the other end of the link, the unit must be used as a "slave" station, the timing of its transmitted pulses being determined by the pulses received from the other end of the link. If this is not done, small differences in pulse-repetition rates would prevent interleaving of the pulses.

The pulse train from the master station is received by the slave station. From the second-detector circuit of the slave receiver, there is filtered out a sine wave having a frequency equal to the pulse-repetition rate. This wave is formed by the pulse generator into time-modulated pulses to be transmitted by the slave station. By proper adjustment of the phasing control, these pulses may be made to occur during the interval when no incoming pulses are due to be received. Even within the limits imposed by this condition, some shifting of the

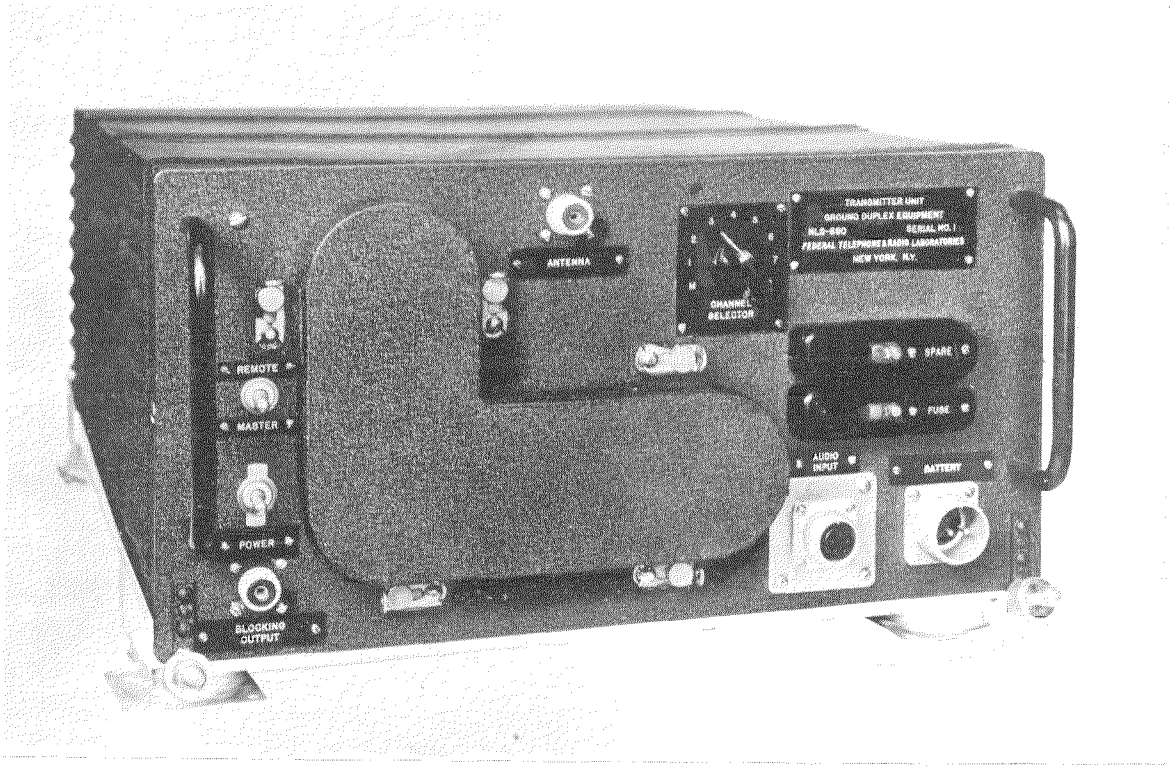

Fig. 3—Experimental pulse-time-modulation transmitter designed to operate with the U. S. Army EE-8 field telephone system.

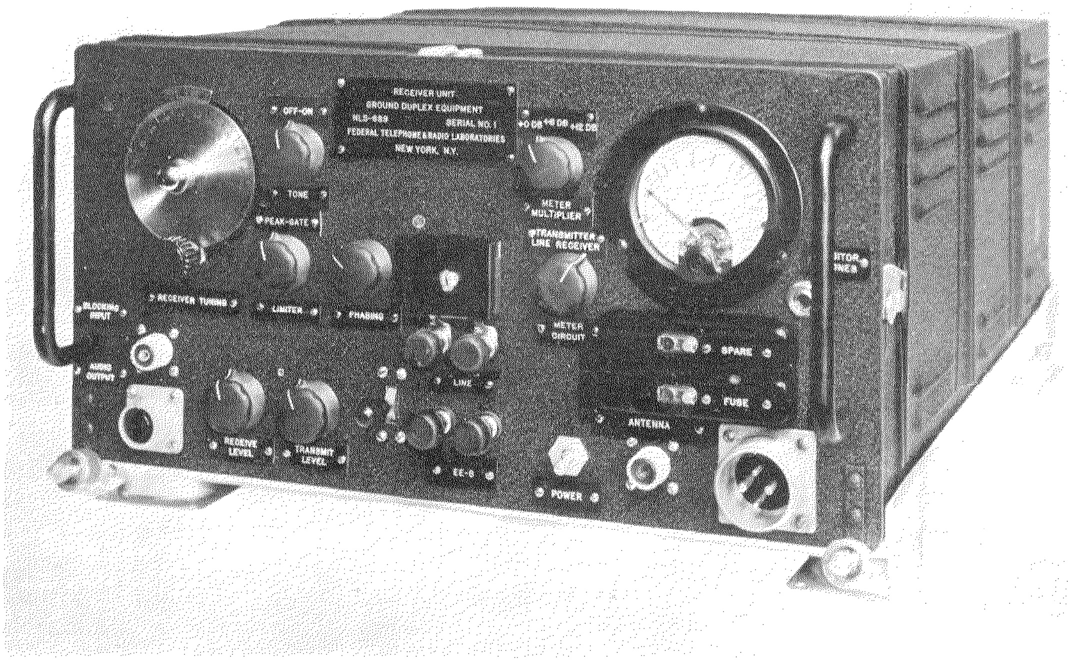

Fig. 4—Experimental receiver to operate with the transmitter shown in Fig. 3.

phase will be required, depending on the time of transmission of the pulses from one end of the link to the other, to insure that at the master station the incoming pulses do not conflict with the transmitted pulses. This adjustment is not difficult and need be performed only once for any given spacing of stations.

The relative timing of pulses and waveforms throughout the system are shown in Fig. 2. This diagram indicates how the duplex operation is accomplished without interference between incoming and outgoing pulses. As the equipment is not transmitting at the same time that it is receiving, it is possible to use a single antenna for both units, even though they may be operating on the same channel.

2. *Equipment*

Two terminal equipments were constructed and tested at some length in the field. The transmitter and receiver are shown in Figs. 3 and 4, respectively. They are of identical size, being 15 inches by $19\frac{1}{2}$ inches by $7\frac{1}{2}$ inches. The transmitter had a peak power output of 400 watts and an average power output of 5 watts. It was used with an antenna consisting of a dipole in front of a screen reflector. The system operated between 225 and 285 megacycles per second.

A voice-frequency ringing converter has been included in the equipment, so that when conventional field telephones, such as the military type EE-8, are connected to the terminals of the equipment, they will operate exactly as though a wire connection had been provided, ringing and talking in the normal way. This operation is possible even though several miles of wire are interposed between the field telephone and the terminals of the radio equipments.

In field tests, excellent operation has been secured over a line-of-sight path of 23 miles and over a somewhat obscured path of 7 miles. Under these conditions, communication has been observed to be equal to, or better than, the standard telephone circuits.

Progress of Telecommunication Services in British Post Office*

1. *Telephone Exchanges*

THE POLICY of the General Post Office is to have automatic working at all exchanges and, before the war, the conversion of existing manual exchanges to automatic working was progressing at a rapid rate. At the outbreak of the war, work on mechanisation had to be curtailed to essentials; in some cases equipment which was being manufactured was completed but stored for installation at the end of hostilities.

During the war, there was a tendency for the number of automatic telephones to decrease because automatic exchanges destroyed by enemy action were replaced by manual exchanges. In 1921, there were 16 automatic exchanges: by 1935, these had grown to 1600, and at the present date there are over 3700 automatic exchanges. Since the end of the war, there has been great difficulty in replacing exhausted exchanges, converting manual exchanges, and extending exchanges that have been overloaded for a long time because of the very great difficulty in getting buildings and equipment.

Priority for building materials and labour is given to the building of new houses and to the repair of buildings damaged during the war. The amount of building which can be provided for telecommunications is only a very small proportion of what is required. Consequently, many expedients have had to be adopted to extend the lives of exchanges and to avoid the need for buildings. At those exchanges where no other method of giving relief is possible, it has been decided that pre-fabricated buildings will be used. The requirements for exchange and subscribers' equipment cannot be fully met by manufacturers because of the competing demands for materials and labour for essential needs, such as export, fuel and power, and railways. The service, however, continues to expand and the rate of installation of subscribers' telephones is now about double the pre-war rate.

2. *Mobile Unit Automatic Exchanges*

Immediately before the war, it was decided to install, experimentally, a few unit automatic exchanges (UAX) on special motor chassis. These were designed for installation at short notice in cases where service could not be maintained at an existing manual or automatic exchange as a result of damage or other unforeseen circumstances. During the war, 14 mobile exchanges were produced, of which 2 were supplied to the War Office, the remainder being kept in reserve by the Post Office for the restoration of exchanges damaged by enemy action. It was found in practice that they were required only on a few occasions for this purpose.

Although the damage aspect has largely disappeared, an increasing use of unit automatic exchanges has been made in cases in which Sub-Postmasters have found it urgently necessary (frequently on grounds of ill health or advancing age) to ask to be relieved of the work of manual exchange operating and alternative arrangements could not immediately be made. The mobile units have also been useful in converting from one type of exchange to another in an existing building. It has recently been decided to increase the number of units and to retain them as standard items.

3. *Improvements in Telephone Switch-Rooms*

3.1 LIGHTING

During the war, many switch-rooms were permanently "blacked out," and, in consequence, much attention has been directed to the question of artificial lighting. A number of these exchanges were provided with fluorescent tube lighting, and, as a result of this experience, a series of experiments with this type of lamp is being carried out at selected exchanges to determine the best lighting arrangements.

3.2 HEIGHT OF SWITCH-BOARD

Standard switch-boards are made in two heights, 4 feet $8\frac{1}{2}$ inches and 6 feet $4\frac{1}{2}$ inches,

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* From material and illustrations supplied by the British Post Office.

Where the vertical multiple is high, the operating effort is increased, and it also tends to give the staff a "shut in" feeling, particularly where the ceiling is low. In present designs, arrangements are being made to use the lower type of switch-board wherever possible.

3.3 COLOUR AND FINISH OF SWITCH-BOARD

The standard finish of polished mahogany has been criticised on two grounds. First, it makes the switch-room dark and gloomy, and second, it can give rise to unwanted reflections. As an experiment, a switch-board is being finished in limed oak, which is light in tone and has a surface that is practically matt.

3.4 KEYSHELF

This is covered with red fibre on present designs; fibre is one of the few materials that will stand up to the wear and tear experienced. An experiment with green fibre is being undertaken, as it is thought that this will have a more pleasing appearance.

3.5 MULTIPLE

The multiple face accommodates strips of jacks and lamps, labels, and plain strips to fill up spare capacity. A number of experiments involving the use of various colours and new materials are being undertaken, and polished surfaces are being replaced by matt finishes where possible.

3.6 FUTURE SWITCH-BOARD DESIGNS

The cordless type of switch-board is being considered for future exchanges. With this equipment, there is no vertical multiple, and the appearance of the positions can be made much cleaner and more attractive. Also, the physical effort of operating is much reduced. In designing switch-boards for the future, efforts are being made to simplify the operator's work, and attention is being given to motion study.

3.7 SWITCH-ROOM LAYOUTS

At present, switch-boards in large exchanges are arranged with their backs to the wall round the sides of the room. With possible future developments such as the cordless switch-board in view, consideration is being given to other arrangements. For instance, switch-boards might be installed back-to-back and in short suites across the width of the room.

4. *Private Branch Exchanges*

During the war, a new type of switch-board was developed for use in large manual installations. This, the P.M.B.X. No. 1A, supersedes the C.B. No. 9. Its principal advantages are that it employs lamp signalling, and it is constructed of standard parts; it is also more compact, and is now standard for installations of over 180 lines.

This new equipment is a 24-volt CB-type multiple switch-board, with capacity for 800 extensions and 160 exchange lines, private circuits, or inter-switch-board lines. It is built up in one-position two-panel sections. The extension multiple is made up of strips of 20 jacks of the break-jack type, and the exchange line multiple is of strips of 10 jacks of the branching type. Both multiples have a four-panel repetition.

There is no separate answering field, calling lamps being fitted in the multiple and associated with selected appearances of the lines to which they relate. Each extension has a single calling lamp, while exchange lines are given a second ancillary lamp at a different multiple appearance.

The design also embodies a "follow-on call trap" which, in cases where an extension-to-exchange connection has not been cleared at the private branch exchange, prevents a fresh incoming call from ringing the extension bell and gives a supervisory flash instead. Also included is a "calling in" facility, whereby an extension user can press a special button to flash the private branch exchange operator during a call to the public exchange without disconnecting the call.

5. *Local Lines*

In the pre-war period, meticulous and exhaustive development studies were made throughout the country in order that the planning of local line networks might be placed on a sound footing. During the war, owing to dispersal of industry, evacuation of the population, war damage, and demands of the armed forces, these studies became more and more out of date. Subsequently, the return of many businesses and evacuees to the towns, closing down of war installations, redistribution of industry, and replanning of industrial development, produced further complications. As a result, the existing development studies are in need of drastic revision, and this work is already in hand. In the meantime, special studies are being made in areas where there is an urgent need of new plant to meet accumulated demands for telephone service. Completely new studies must be made in certain towns, e.g., Liverpool, where records were destroyed by bombing, and joint engineering and commercial surveys are being made in this connection.

Good progress is being made with duct- and cable-laying schemes, and the provision of auxiliary joints and cross-connection frames in street cabinets and pillars will ensure the flexibility that is so necessary during periods of unforeseen development. As a result of improvement in telephone instruments, it may be possible to use lighter-gauge conductors than hitherto and effect corresponding economies in local cable schemes.

In country districts, new distribution points are being created and new lines provided at an unprecedented rate. Pole shortage and the uncertainty of future supplies of timber suitable for poles have compelled the recovery of poles from abandoned routes. The use, in spite of higher costs and the need for special fittings, of non-wooden poles is being considered.

6. *Rural Areas*

The policy of the Post Office is to provide service in country districts by means of small, unattended, automatic exchanges, and to install such an exchange in any area where a minimum of eight subscribers is forthcoming. This policy involves the conversion of all small manual exchanges in rural areas to automatic working, and has also led to the opening of a very large number of new exchanges. Between the wars, the number of country exchanges was increased from about 1600 to about 4000.

The programme of automatisation in rural areas was largely suspended during the war, but will be resumed as soon as conditions permit. Until adequate supplies of equipment are available, however, it will be necessary to concentrate largely on the extension of existing automatic exchanges rather than on work of conversion of manual exchanges.

Telephone service at standard rentals is available to anyone within three miles of an exchange, an additional charge being made where this distance is exceeded. The extent of the coverage is now such that there are comparatively few localities that are more than three miles from an exchange.

7. *Public Call Offices*

It is the policy to provide a kiosk in every village where there is a post office. While much work under this policy has already been carried out, it was almost entirely suspended during the war. It is now being resumed.

In villages where there is no post office, a kiosk is provided on payment by the local authority of £4 a year for five years, the pre-war rental of a private subscriber's line in the provinces for five years. This covers about 10 per cent of the cost, the remainder being borne by the Post Office; at the end of the five years, the full cost falls on the Post Office.

Between the wars, the number of public call offices in rural areas increased from 4700 to nearly four times this figure.

8. *Local Exchange Service*

In London and the four largest provincial cities, Birmingham, Glasgow, Liverpool, and Manchester, the director-type automatic equipment is in use although there are numbers of manual exchanges still to be converted. The London system consists of an area enclosed by a circle of 12½-mile radius round Oxford Circus, and subscribers connected to all but the very earliest type of director exchange can dial directly to all other exchanges in the system. The call fees are recorded automatically.

In the provincial director areas, the ultimate size of the systems will be a circle of 9½-mile radius, although at present there are only 3 director exchanges beyond the original boundary of the systems, which was a circle of 7-mile radius. There is multi-metering (i.e., automatic connection and recording of calls with charges up to 4d.) to all exchanges in the provincial director systems, except from a few of the oldest type of exchanges in Birmingham and Manchester. The equipment programme includes the conversion of existing manual exchanges in the director systems and the introduction of multi-metering at some of the older exchanges where it was not originally provided.

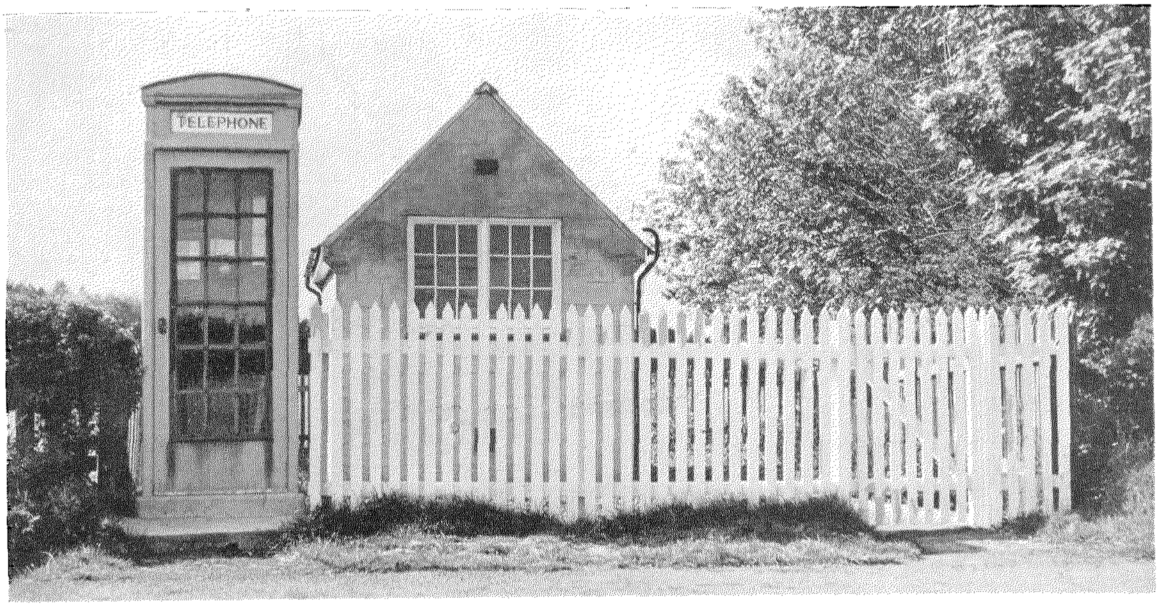
Plans are now being made to extend the range of direct dialling by subscribers on director exchanges to other exchanges within 20 miles of Oxford Circus, and 17 miles of the centre of provincial director systems. Development work is also being carried out to give multi-metering from automatic exchanges adjacent to a director system to exchanges within the director system; these facilities will be introduced first in the London area. It has been decided to introduce director working in Edinburgh and one director exchange, Craiglockhart, was opened in 1946. The size of the director automatic area in Edinburgh will be a circle of 5-mile radius.

Multi-metering facilities are now given at the largest (director) exchanges and at the smallest (unit automatic) exchanges. Multi-metering is a standard facility for unit automatic exchanges although access to every exchange within range is not always possible because tandem equipment is not available at intermediate exchanges.

The equipment at the medium-sized automatic exchanges and many of the larger exchanges that are not big enough for director working is of the non-director type. At these exchanges, subscribers at present can dial only to exchanges up to 5-miles chargeable distance. The relay set, which is used to give the appropriate metering, i.e., 1d. on routes up to 5 miles, has been modified to give second, third, and fourth-fee metering so that multi-metering can be given on calls to directly connected exchanges up to 15-miles chargeable distance.

It is estimated that the tandem traffic from non-director exchanges is small, that it is generally about 5 per cent of the total untimed junction traffic and rarely rises above 10 per cent. This tandem traffic could be completed automatically by using equipment that would correctly route and register traffic passing over one route. The equipment is relatively expensive and not easy to maintain. It is probable that the automatic completion of the small amount of tandem traffic will not prove to be justified economically.

The development of the relay set to give multi-metering on directly connected routes has recently been completed, and data are now being finished for the design of equipment to provide multi-metering facilities at several new and existing exchanges. For the present, it is not proposed to make any arrangements for the multi-metering of the non-director tandem traffic.



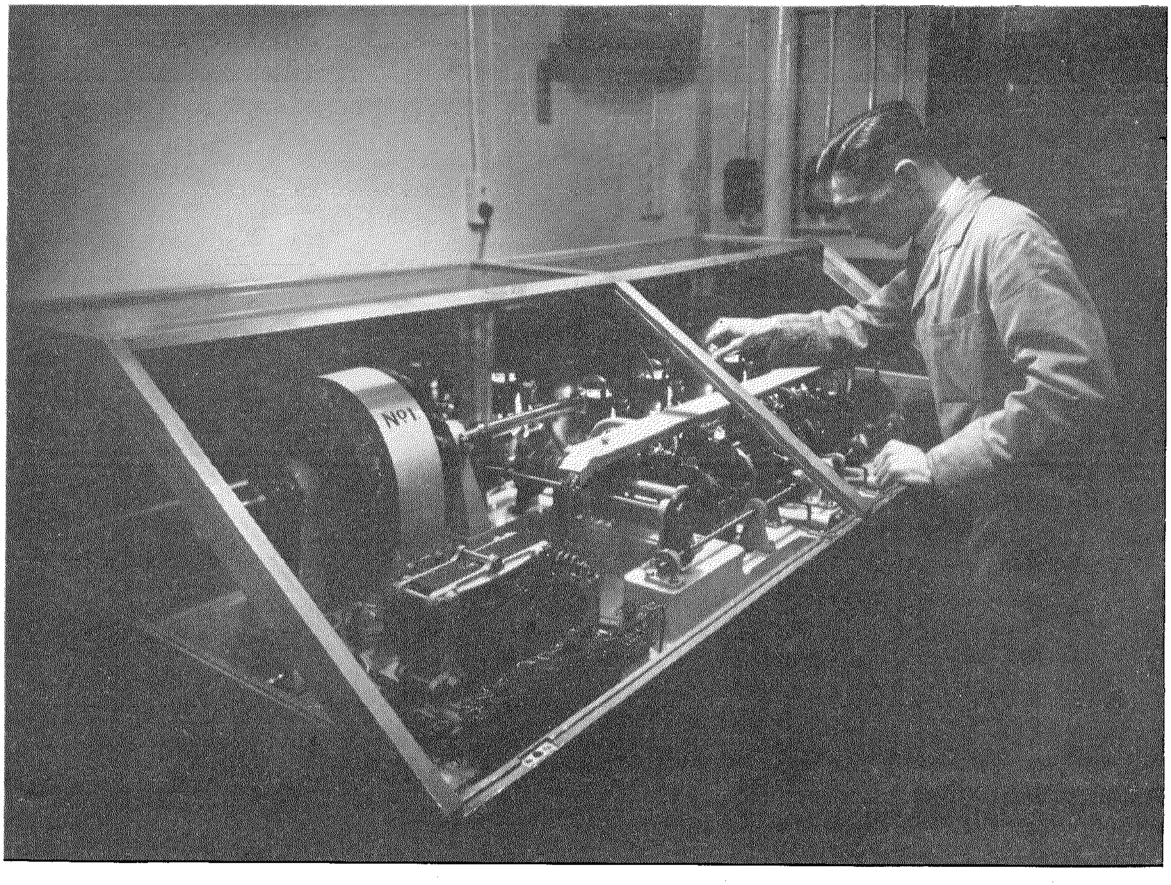
Unattended automatic exchange and kiosk at Beal, Northumberland.

9. Provision of Telephone Service

During the war, the Post Office suspended the provision of all local subscribers' cables other than those immediately required for war or war production purposes, and the same policy was followed in regard to extending telephone exchanges. The Post Office also released 16,000 of its 41,000 engineers for service with the Forces. It was, therefore, impossible to meet many applications from the public for telephone service and, when the war ended, the outstanding applications amounted to 250,000. The difficulty of dealing with this accumulation of orders was added to by the receipt of new orders at an unprecedented rate. During 1946, the average demand was for 70,000 telephones per month, which is more than double the prewar figure. A peak was reached in October, 1946, when the demand reached 83,000, but it has now fallen to a fairly steady rate around 45,000 per month, which is about 50 per cent higher than before the war.

The rate of installation of new telephones was stepped up considerably in 1946 as engineering staff returned and reached 70,000 per month, but the effect of this rapid rate of connection was to use up reserves of exchange equipment and spare wire in underground cable. The amount of engineering work and materials required to connect a given number of subscribers has steadily increased, and it has not been possible to maintain the high rates achieved last year. The present rate of installation is about 53,000 per month as compared with 31,000 before the war.

Despite the efforts which have been made, the outstanding applications for exchange lines now amount to 450,000. The spare exchange equipment is exhausted at about a fifth of the exchanges and there are no spare wires in the under-



The speaking clock shown above announces the time within an accuracy of 0.1 second. It may be dialled directly in certain exchanges and obtained through operator's connections in some others.

ground cables serving a third of the distribution points. In this country therefore, as in America, the provision of telephone service to many ordinary applicants is subject to appreciable delay. It is still necessary to impose limitations on the amount of constructional work undertaken to connect an individual subscriber, and a system of priorities is in operation to ensure that telephone service is provided first to those with a special claim in the public interest.

The Post Office has an ambitious programme for the extension or replacement of telephone exchanges and the expansion of the underground plant but the fulfilment of these plans is being seriously handicapped by inadequate supplies of stores and equipment, due mainly to shortages of raw materials.

10. Emergency Call Service

This service is provided to enable a telephone user on an automatic exchange to obtain, in an emergency, by dialling "999", an immediate reply from the auto-manual board operator who is able to connect to the police, fire, or any other emergency authority without delay. No coins are required to obtain such a call from a coin-box telephone. The service was available at London and Glasgow before the war. Since the war, it has been installed in all but the smaller automatic exchanges.

11. Speaking Clock

This service is available to subscribers on automatic exchanges in London and 13 other towns and cities, and on the manual exchanges in London, Birmingham, Glasgow, and Manchester. To obtain the time, subscribers dial TIM in director areas and "952" in non-director areas. They are connected automatically to a speaking record of the time which is accurate to 1/10 of a second. Calls to the speaking clock cannot be obtained automatically from call offices but are obtained by calling the operator who connects the calls manually.

12. Long-Distance Telephone Service

During the war, the provision of circuits for the public service did not keep pace with the increase in traffic, as a very large proportion of the new plant laid down by the Post Office during the war years was used to provide private wires to meet the needs of the defence forces and the armed forces of the allies stationed in Great Britain and Northern Ireland; also, the Post Office suffered a heavy loss of trained technical staff to the signals branches of the navy, army and air force. Although some temporary staff, including women, was recruited to take their places, the numbers were fewer and, of course, they were less skilled and lacked the experience of the regular staff. There was, in consequence, a deterioration in the quality of the long-distance service and also in the maintenance of the plant, which in turn added to the service troubles resulting from bombing, congestion of traffic, etc.

Despite these difficulties, traffic is now about 80 per cent higher than when the war began, the number of long-distance calls completed each week being almost 4½ million compared with about 2½ million in the corresponding period of 1939. With the very active state of business throughout the country, traffic is still growing. Doubtless the war-time dispersal of factories and the evacuation of populations from towns to less vulnerable districts, with the consequent separation of families, has fostered the "telephone habit." Moreover, despite a 50 per cent increase of the trunk-call charges, the service is still inexpensive, the highest rate being 3s. 9d. for 3 minutes for any distance of more than 125 miles.

Since D-Day, large numbers of the private wires provided for the defence and armed forces have been relinquished and those found suitable have been added to the public system.

The number of circuits of over 25-miles chargeable distance now (July, 1947) totals about 13,600 compared with 6775 in April, 1939, an increase of about 100 per cent. This increase in the number of circuits, together with an easing in the staff situation, has resulted in a rapid improvement in the service and, except for a few routes on which a shortage of circuits still exists, the pre-war speed of connection has been largely restored. There is still a shortage of experienced operating staff, however, particularly in London, where the speed of answer, though steadily improving, is not yet back to the pre-war standard.

Before the war, the long-distance service was designed to enable 90 per cent of calls made during the day-time to be completed on demand, and it is now the aim to provide sufficient circuits to enable 98 per cent of calls to be completed on demand. For the provision of circuits necessary to meet this improved standard, and to cater for general growth in the volume of traffic, plans have been made for the laying of an extensive network of coaxial cables linking all the main towns. Many of these coaxial cables will include up to 400 audio-frequency pairs to cater for toll circuit requirements, and by this means the laying of separate audio-frequency cables will be avoided and full use made of all available duct space.

13. Trunk Service—Direct Dialling

It is possible for operators in trunk exchanges to dial directly to subscribers at distant centres by transmitting over the trunk lines dialled impulses of direct current or of alternating currents of certain frequencies within the frequency range of the human voice. This is already employed on a large number of short- and long-distance routes, and the scope of such systems is being progressively extended so that within a number of years trunk operators will be able to call most subscribers within the United Kingdom simply by dialling and without the need for another operator.

It is not proposed, at least in the immediate future, to arrange for direct dialling of trunk calls by subscribers.

14. Mechanical Trunk Fee Accounting

Recently, methods of mechanising the accounting for subscribers' trunk calls have been investigated with a view to eliminating the present tedious and laborious manual sorting of tickets on which calls are recorded. The proposals at present being considered would involve the replacement of the present paper tickets by cards that would be perforated, after leaving the exchange, with holes corresponding to the manuscript data recorded thereon. The tickets would then be sorted at high speed by machines into subscribers' number or any other order necessitated by the accounting processes, e.g., tickets from non-consecutive private-branch-exchange lines would be picked out for association with the tickets for the main number. It is probable that the pricing of trunk tickets would be performed mechanically if the schemes under consideration were adopted. After punching and sorting, the cards would pass through a tabulator, which is a machine capable of listing all the particulars on the cards, carrying totals, and printing these as required. Thus the subscriber's trunk statement would contain an abstract of the data recorded on the tickets, as much or as little information as required being supplied. It is, in particular, desired to furnish subscribers with the name of the exchange called on trunk calls.

There are hundreds of millions of trunk tickets handled every year, and it is hoped that considerable staff savings would be achieved from mechanising the accounting processes. As an offset, more expensive machinery would be required, but it is hoped that overall savings would result. It should be fairly easy to extend the mechanical system to deal with rentals and local calls, if this were found to be useful and justifiable.

15. Census of Long-Distance Telephone Traffic

In October, 1945, the British Post Office took a census of long-distance telephone traffic to obtain information on which to base post-war traffic and cable network planning. Full details from all tickets prepared on three normal days were entered on cards which were subsequently perforated for analysis by the punched-card system. Some 800,000 cards were mechanically sorted to provide a record of calls from each trunk centre to every other trunk centre during each half-hourly period throughout the day, thus enabling a more accurate estimate of relative degrees of community of interest between trunk centres to be made. The addition to the cards of perforations corresponding to the standard routing of the calls made possible a further analysis showing the amounts of traffic proper to each existing route.

The volume of traffic in each charge step and the revenue obtained from this traffic has been analysed, but this leaves a considerable amount of data which can still be extracted from the census figures including, to quote one example, the proportions of the various types of call, e.g., personal, transferred charge, and fixed time, made from each centre over various distances and the times at which the demand for these services reaches a peak.

16. Trunk Network

During the war, the British trunk network was rapidly expanded as a result of defence requirements, and 12-channel carrier now forms the backbone of the system, audio-frequency cables being used for shorter junction routes. The 12-channel standard units, with appropriate modulating equipment, may be worked over (A) a paired cable of 14 or 24 pairs in which each pair may carry up to 24 channels and separate cables are used for "go" and "return," (B) an air-dielectric coaxial cable, in which each tube carries up to 600 channels in one direction, or (C) a solid-dielectric coaxial submarine cable, in which the number of groups is limited by the attenuation of the cable between adjacent repeaters.

17. Radio Links—Development for Public Traffic

In the past, extensive use has been made of ultra-short waves (60 megacycles per second and above) for radiotelephone communication over sea routes up to 100 miles. Many of these have been multichannel amplitude-modulated services. In some cases, the radio link has served to carry the 12 channels of a line carrier system without demodulation to audio frequencies at the radio terminals. A development in recent years has been to apply frequency modulation to these 12-channel radio systems with resulting improvement in intermodulation characteristics and in signal-to-noise ratio. A number of such systems are in operation or are planned.

18. Inland Telegraph Service

Just before the war and after an exhaustive study of the traffic characteristics of the British

Teleprinter switch-room in Manchester.

telegraph service, it was planned to install an automatic teleprinter switching system by which any telegraph office equipped with teleprinters could communicate directly with any other teleprinter office. War broke out before the first installation was started, and the heavy commitments of work for the armed forces prevented the scheme being carried out. This resulted in the retention of the existing point-to-point system which, by reason of its heavy concentrations of traffic at the large towns, was particularly vulnerable to air attack. To reduce these concentrations as far as possible, a policy of dispersal was adopted by which traffic circulated through the smaller towns, but which, in consequence, increased the number of retransmissions. To counter this effect, which was aggravated by a substantial increase in traffic, a manual switching scheme was planned and installation of the first two switch-boards began in 1943. This scheme has developed steadily since that date, and six switching centres are now installed and working. Approximately 150 offices will be connected to the system and something like 100,000 telegrams per day signalled over it when it is complete.

In February, 1945, when it was evident that the end of the war was not far off, steps were taken to begin the preliminary work for recommencement of the automatic switching programme. Much valuable knowledge of telegraph switching schemes had come from the war-time experience and the pre-war design was entirely re-examined and improvements made. In particular, it has been decided to abandon the use of the standard Post Office No. 3 teleprinter and to adopt one using the International Teleprinter Alphabet No. 2. The commencement of the scheme is planned for 1949, and it is hoped to complete it by 1954. Ultimately, approximately 700 teleprinter offices will be connected and will be given full inter-communication by dialling. There will be 26 automatic switching centres. The system will incorporate certain new features such as automatic return of the answer-back signal and suspense conditions on calls to engaged lines.

A system of automatic distribution of calls from telephone subscribers wishing to dictate telegrams is also being developed. The principles are similar to those used in such systems in other countries, that is, distribution to disengaged operators, and storage and queueing when no disengaged operator is available. The usual waiting call indicators will be provided.

19. European Telegraph Services

Voice-frequency telegraph systems on coaxial submarine telephone cables have been set up and improved performance and a considerable increase in the number of telegraph channels to the continent have resulted. The additional channels provide a sufficient margin to permit the establishment of private telegraph circuits for the larger users between Great Britain and the continent, and also for the development of Telex services with the continent, thereby catering for short-duration connections between teleprinter renters in Great Britain and the continental countries. Telex service with Belgium, Holland, and France has been established, and it is hoped to open service to Switzerland shortly.

20. Extra-European Telegraph Services

The public telegraph services between the United Kingdom and places outside Europe are operated by telegraph companies, and not by the British Post Office. Telegrams for all parts of the world are, however, accepted at all postal telegraph offices.

21. Development of Overseas Telephone Services

During the war, international telephone communication was almost at a standstill. All land-line communication with the continent of Europe had ceased by the time France was overrun; the radiotelephone services were maintained only when absolutely necessary and then under severe censorship restrictions.

From D-Day onwards, telephone and teleprinter circuits have been progressively restored, first for the use of the armed forces for communication between bases in England and advanced positions on the continent, and later for public service.

Plans for post-war development are based on carrier technique and take advantage of the lessons learned in the manufacture, laying, and use of the carrier cables across the English Channel during the final phases of the war.

The majority of the cross-channel cables that were cut during the war had been designed for audio-frequency use although a few had been designed for carrier working; the full facilities amounted to about two hundred circuits. Circuits in the war-time carrier cables have been set up as public-service circuits and work in the manufacture and laying of new coaxial carrier cables is well in hand to meet future needs.

To cater for Anglo-French requirements, two of the cables laid during the war provide 60 channels per cable and, when the 1939 St. Margarets Bay—Calais carrier cable is repaired, an additional 168 circuits will become available.

Traffic between Britain and Belgium is already considerably in excess of the pre-war level. Many additional circuits will be required, not only to meet the needs of Belgium, but also of services beyond. The provision of a new single-core coaxial cable of a larger and improved type is in hand and will yield over 200 circuits.

To extend the additional French and Belgian circuits from St. Margarets Bay to London and to cater for possible television requirements, a four-tube coaxial cable is to be laid, which will be able to provide for up to 1200 circuits. Simultaneously, the French and Belgian administrations are making plans for extending the circuits across their countries.

A new cable is already in course of manufacture for the Anglo-Dutch route which, it is hoped, will be ready for service by the end of 1947. The cable, which will be 86 nautical miles in length, will be of the coaxial type similar to that which is proposed for the new Anglo-Belgian cable. The possibility of fitting ultimately two submerged repeaters to make the cable capable of providing 250 circuits is being considered.

Thus the British Post Office, in conjunction with the other administrations concerned, is looking well ahead. As a member of the Comité Consultatif International Téléphonique, Britain is participating in a study of the problem of providing an "on demand" service in Europe, and is planning its communications accordingly.

On the overseas radiotelephony side, the services have been extended to meet the growing demand, and additional services are opened as and when the need arises. The Rugby transmitting station is to be extended, and a new receiving station will shortly be provided. Single-sideband equipment is available for all services and will be brought into use generally as equipment becomes available at the overseas terminals.

22. *Ship-Shore Radiotelephone Services*

The pre-war facility of radiotelephone communication between ship and shore with a link service to the inland telephone system has been restored and expanded. In addition, consideration is being given to the provision of a very-high-frequency radiotelephone service for direct two-way communication between ships and shore to facilitate navigation, docking, etc. in coastal, port-approach, and harbour areas.

SOJ-12 Open-Wire Carrier Telephone Systems in South Africa

By D. P. J. RETIEF

Department of Posts and Telegraphs, Pretoria, Union of South Africa

and

H. J. BARKER

Standard Telephones and Cables, Limited, London, England

AN *SOJ*-12 SYSTEM provides 12 additional two-way speech channels over an open-wire pair on which a three-channel carrier telephone system and the usual voice facilities may already be operating. That is, it increases the capacity of a pair from 4 to 16 telephone channels.

With the ever-increasing demand for long-distance communication facilities in the Union of South Africa, the administration has embarked on a comprehensive programme of installing such systems. The initial programme includes 12 systems, 6 of which are long-haul circuits utilizing one or more repeaters, the other 6 being for short distances, directly connecting nearby communication centres without intervening repeaters. A rough map of these routes is shown in Fig. 1. It is proposed to install 20 additional systems by the end of 1948.

The carrying-out of such a programme involves considerable work on the open-wire lines. The administration has reconstructed and re-transposed all existing routes concerned, and has on hand the building of some 2000 to 3000 miles

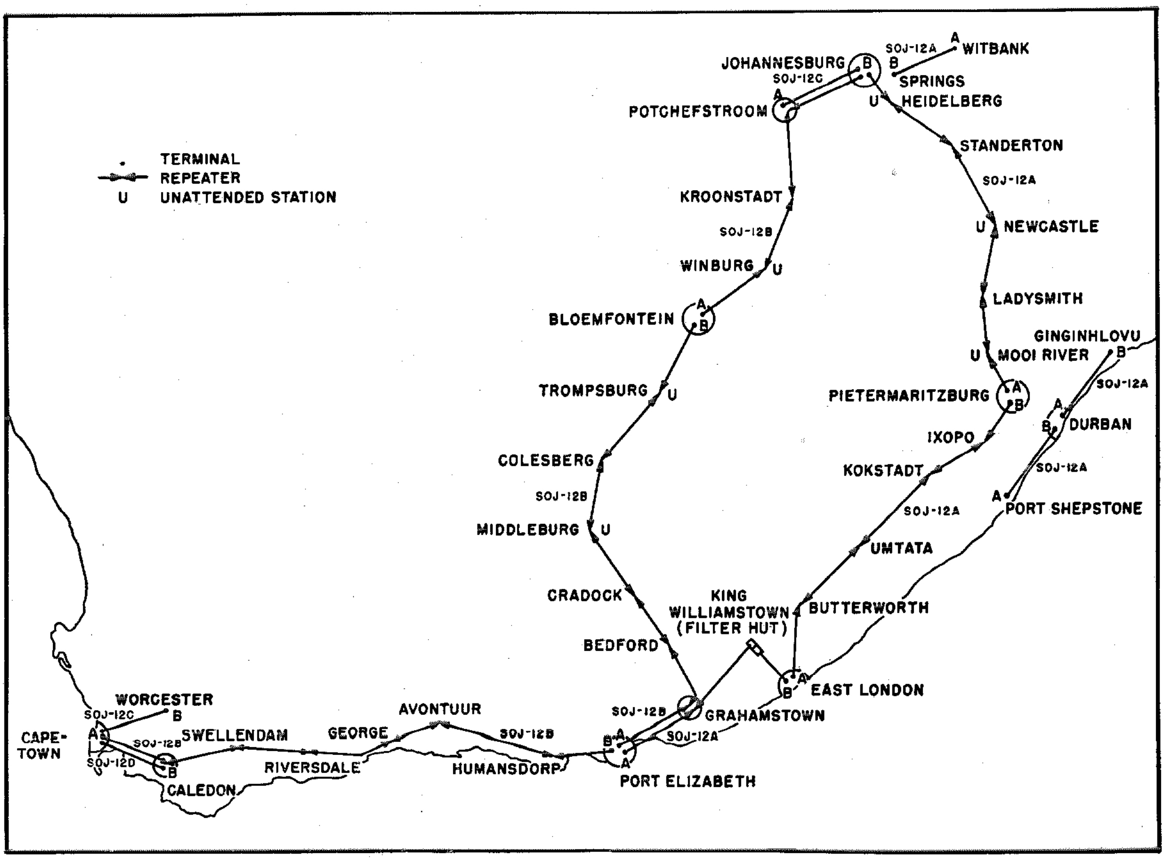


Fig. 1—Map of the initial 12 *SOJ* systems installed in the Union of South Africa.

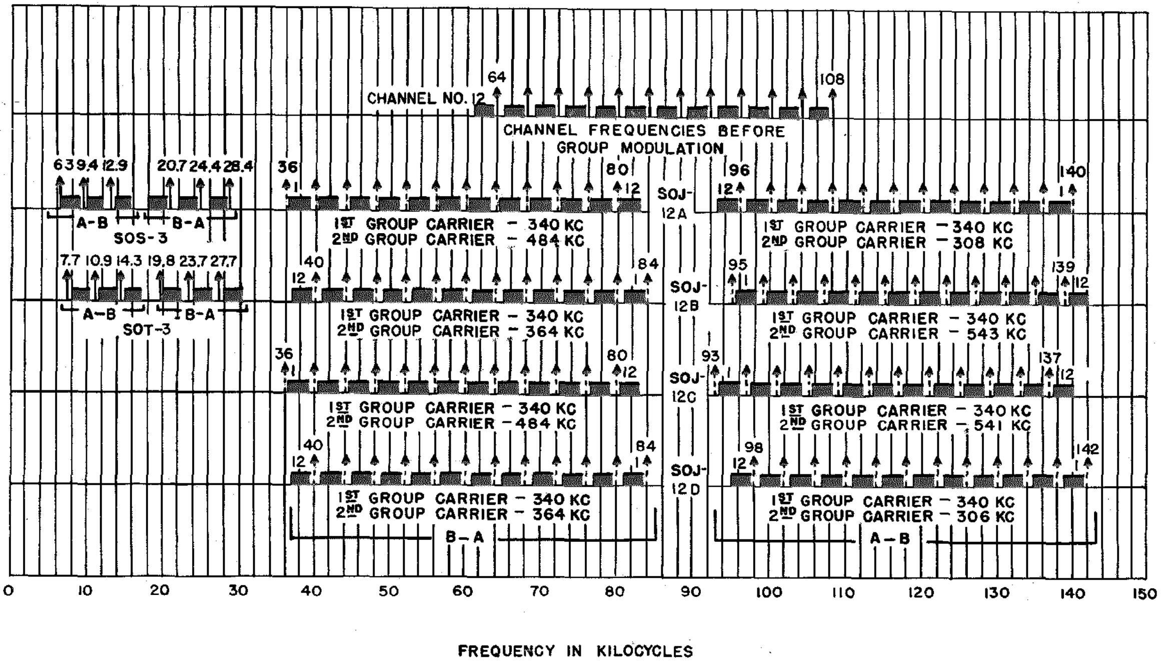


Fig. 2—Frequency pattern of the four variations of the *SOJ-12* system. Real carriers are shown by solid arrows and virtual carriers are indicated by dashed arrows. The *SOS-3* and *SOT-3* systems have been included to show how the *SOJ-12* may be added to circuits already using these 3-channel equipments.

of new *J*-type trunk routes, suitable practices being followed to enable further *SOJ* systems to be added at a later date, up to a total of from 8 to 12 systems.

It is the purpose of this article to describe the application on the Johannesburg-Bloemfontein route of the first *SOJ* system to be installed in the field.

1. *SOJ System*

The system operates on a "grouped-frequency" basis, i.e., different frequency groups are used in the two directions of transmission on the line, equivalent to 4-wire operation over the pair. Four variants of the system are available, designated *SOJ-12A-D*, utilizing frequency inversion and staggering to ease the problems of transposition design and crosstalk whenever more than one system is required to operate on a route. Similar principles are already widely used in the application of three-channel carrier systems. The frequency allocations are shown in Fig. 2.

Figs. 3 and 4 show the block schematic arrangements of typical terminal and repeater stations. Outgoing speech currents on each of the 12 channels pass to the copper-oxide modulators and crystal band filters and then to a common transmitting group circuit. Interposed between the modulators and 4-wire terminating sets are 2-wire to 4-wire switching circuits, which provide the facility for extending each channel on a 4-wire basis as an alternative to the 2-wire termination.

As shown in Fig. 5, the frequency range occupied by the channels at this stage extends from 60 to 108 kilocycles, the lower side-bands of the 12 carrier frequencies, 64, 68, ..., 108 kilocycles, being selected by filters. The equipment providing this 'basic' group of 12 channels is common to *SOJ*-12, coaxial, and 12-channel cable systems.

The group of 12 channels passes through a filter, which further suppresses the channel carrier frequencies, to a hybrid coil where two pilot frequencies are introduced and thence to the first group modulator to which a carrier frequency of 340 kilocycles is applied. The upper side-band, occupying the range from 400 to 448 kilocycles, is selected by a filter, and the signals are raised by an amplifier to a suitable level for application to the second group modulator. The

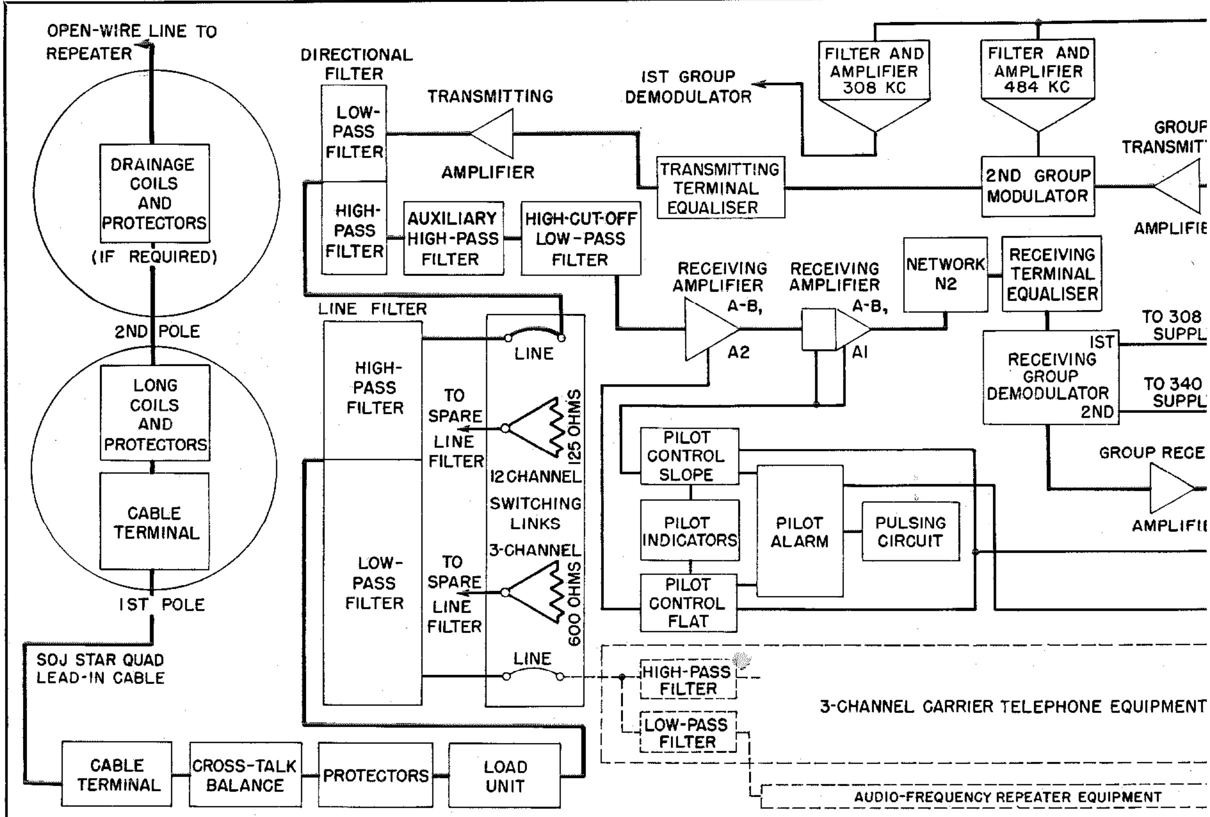


Fig. 3—Schematic of a "*B*" terminal of a typical *SOJ*-12*A* equipment. No filter hut is shown.

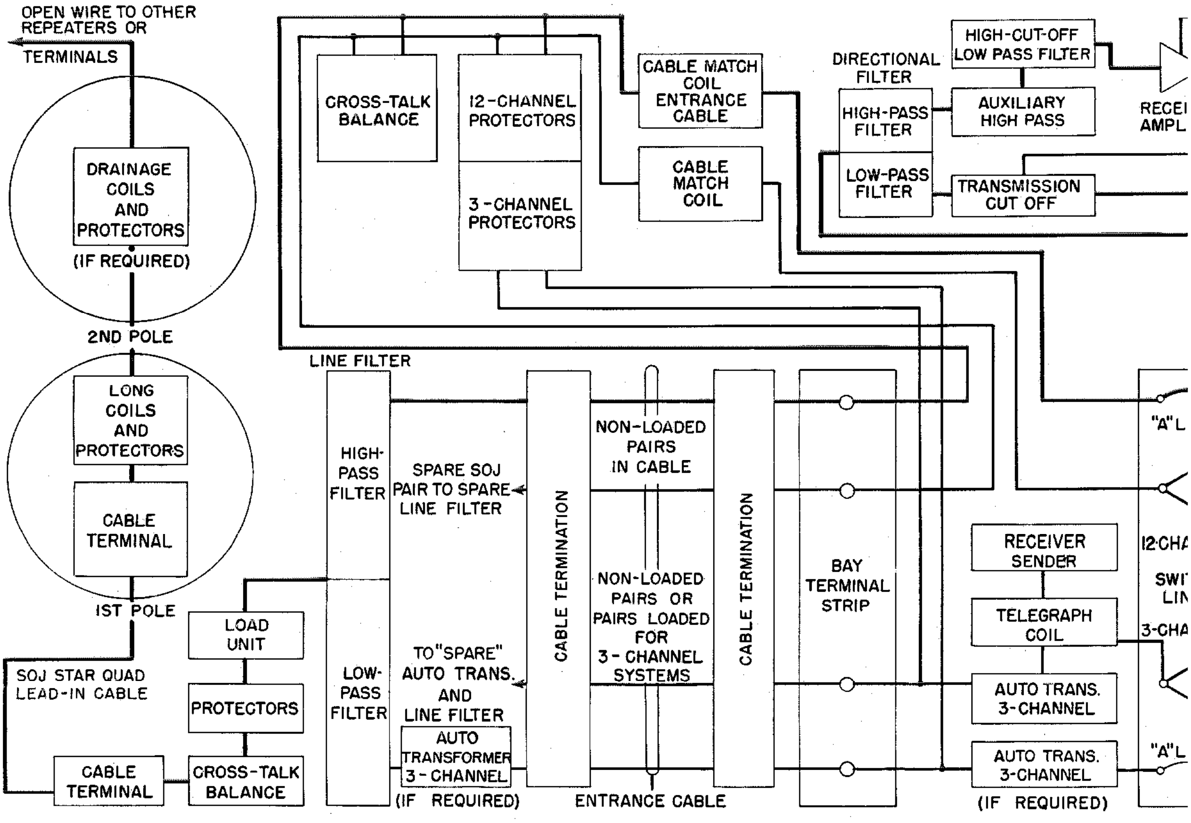
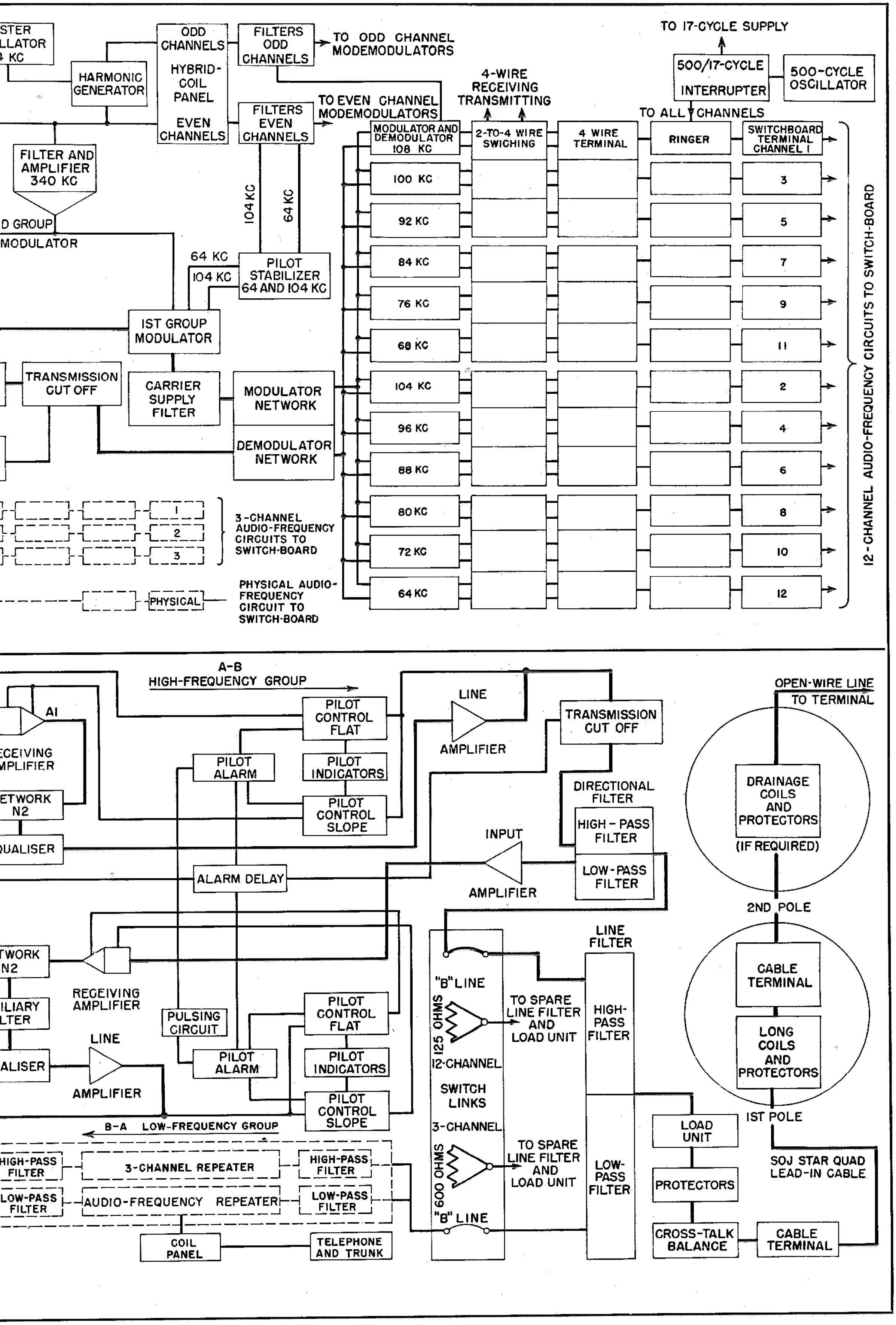


Fig. 4—Schematic of attended repeater station, which controls one unattended repeater. A filter hut contains equipment between the 1st pole and the entrance cable.


STER
LLATOR
KC
HARMONIC GENERATOR
ODD CHANNELS
HYBRID-COIL PANEL
EVEN CHANNELS
FILTERS ODD CHANNELS
TO ODD CHANNEL MODEMODULATORS
FILTERS EVEN CHANNELS
TO EVEN CHANNEL MODEMODULATORS
TO 17-CYCLE SUPPLY
500/17-CYCLE INTERRUPTER
500-CYCLE OSCILLATOR
TO ALL CHANNELS
4-WIRE RECEIVING TRANSMITTING
FILTER AND AMPLIFIER 340 KC
D GROUP
MODULATOR
104 KC
64 KC
64 KC
104 KC
PILOT STABILIZER 64 AND 104 KC
IST GROUP MODULATOR
TRANSMISSION CUT OFF
CARRIER SUPPLY FILTER
MODULATOR NETWORK
DEMODULATOR NETWORK
MODULATOR AND DEMODULATOR 108 KC
100 KC
92 KC
84 KC
76 KC
68 KC
104 KC
96 KC
88 KC
80 KC
72 KC
64 KC
2-TO-4 WIRE SWICHING
4 WIRE TERMINAL
RINGER
SWITCHBOARD TERMINAL CHANNEL 1
3
5
7
9
11
2
4
6
8
10
12
12-CHANNEL AUDIO-FREQUENCY CIRCUITS TO SWITCH-BOARD
1
2
3
3-CHANNEL AUDIO-FREQUENCY CIRCUITS TO SWITCH-BOARD
PHYSICAL
PHYSICAL AUDIO-FREQUENCY CIRCUIT TO SWITCH-BOARD
A-B
HIGH-FREQUENCY GROUP
A1
ECEIVING
MPLIFIER
ETWORK
N2
UALISER
PILOT CONTROL FLAT
PILOT ALARM
PILOT INDICATORS
PILOT CONTROL SLOPE
ALARM DELAY
LINE
AMPLIFIER
TRANSMISSION CUT OFF
DIRECTIONAL FILTER
HIGH-PASS FILTER
LOW-PASS FILTER
INPUT
AMPLIFIER
OPEN-WIRE LINE TO TERMINAL
DRAINAGE COILS AND PROTECTORS
(IF REQUIRED)
2ND POLE
TWORK
N2
RECEIVING AMPLIFIER
ILIARY
LTER
ALISER
LINE
AMPLIFIER
PULSING CIRCUIT
PILOT ALARM
PILOT CONTROL FLAT
PILOT INDICATORS
PILOT CONTROL SLOPE
"B" LINE
125 OHMS
12-CHANNEL
SWITCH LINKS
3-CHANNEL
600 OHMS
"B" LINE
TO SPARE LINE FILTER AND LOAD UNIT
TO SPARE LINE FILTER AND LOAD UNIT
LINE FILTER
HIGH-PASS FILTER
LOW-PASS FILTER
CABLE TERMINAL
LONG COILS AND PROTECTORS
IST POLE
LOAD UNIT
PROTECTORS
CROSS-TALK BALANCE
CABLE TERMINAL
SOJ STAR QUAD LEAD-IN CABLE
B-A LOW-FREQUENCY GROUP
HIGH-PASS FILTER
3-CHANNEL REPEATER
HIGH-PASS FILTER
LOW-PASS FILTER
AUDIO-FREQUENCY REPEATER
LOW-PASS FILTER
COIL PANEL
TELEPHONE AND TRUNK

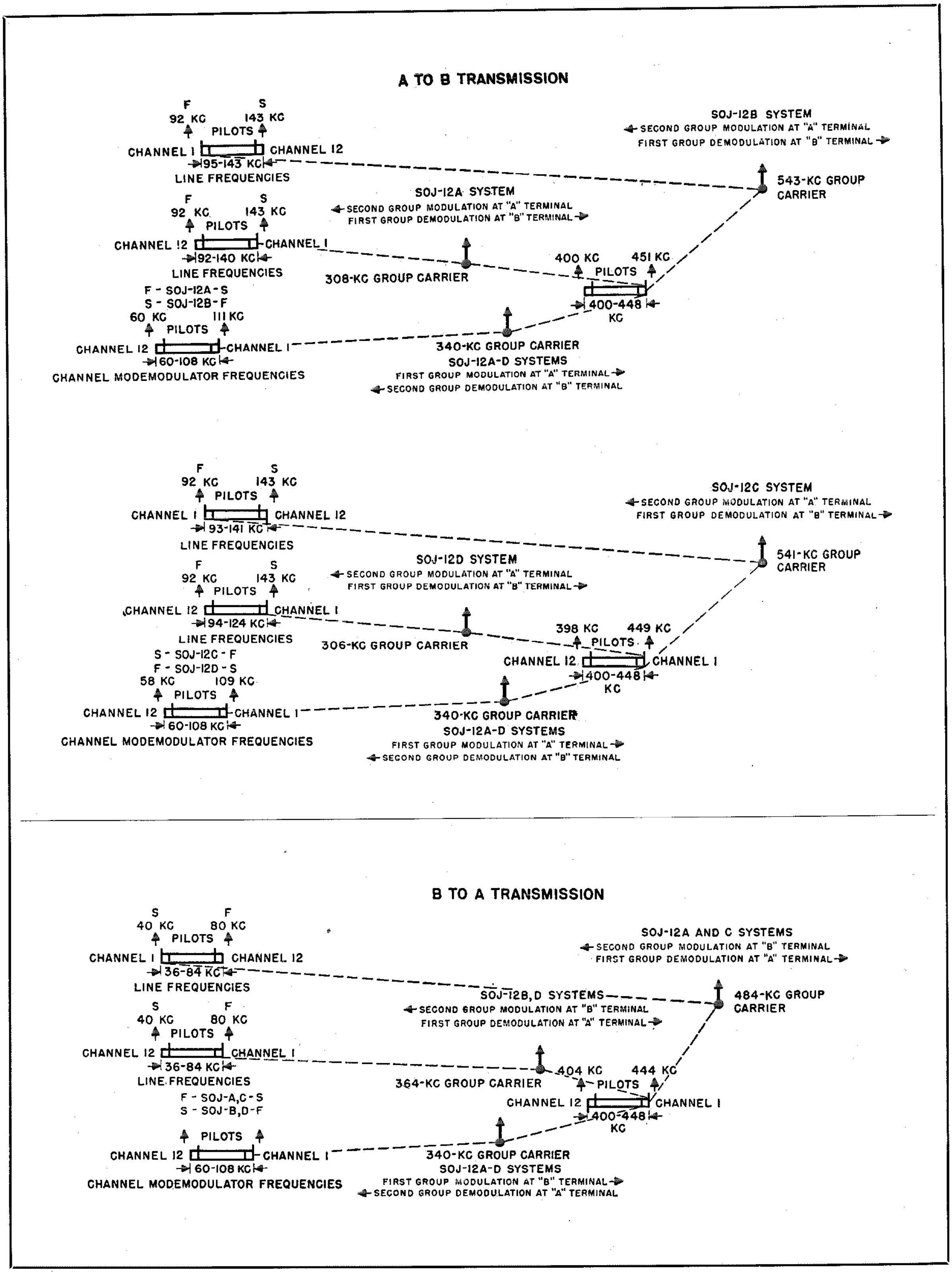


Fig. 5—Group-frequency translations of *SOJ*-12*A*-D systems. *F* denotes flat- and *S*, slope-gain-control pilot.

frequency used for the second stage of modulation is dependent on the direction of transmission and type of system, being 364 kilocycles at Johannesburg and 543 kilocycles at Bloemfontein. The output passes through a low-pass filter, which selects the lower side-band of this modulation process. Compensation for the distortion introduced by the various items of equipment is provided in the equaliser to give a nominally flat characteristic over the required line frequency range, i.e., 36 to 84 kilocycles in the direction from Johannesburg to Bloemfontein and 92 to 143 kilocycles for the reverse direction.

The transmitting amplifier, which is designed to combine high gain with large power-handling capacity, raises the signals to a level suitable for transmission to line (+17 decibels relative to the transmitting trunk switch-board per channel). The signals pass to line by way of the directional and line filters and line plant.

On the receive side, incoming signals pass through the regulating equipment, which is described later, to the demodulation circuits, which function in a similar way to the modulating circuits on the transmit side, to translate the incoming line frequencies to the basic group from 60 to 108 kilocycles prior to the final demodulation process for separating the 12 channels.

The carrier supplies are derived from a 4-kilocycle valve-maintained tuning-fork oscillator, which operates into a harmonic generator. These oscillators have a high degree of frequency stability, being maintained readily in practice within 18 parts in a million of the nominal frequency. Odd harmonics of 4 kilocycles are produced in a saturated coil and even harmonics in a copper-oxide rectifier. The frequencies required are selected by crystal filters, followed in the case of the group frequencies by amplifiers to raise the carrier to a level suitable for application to the group modulators.

Three of the group frequencies, namely, 306, 541, and 543 kilocycles, which are required for certain frequency allocations, are not multiples of 4 kilocycles. For their derivation, a 5-kilocycle oscillator is locked to the main 4-kilocycle supply and its output is used to modulate an appropriate harmonic of 4 kilocycles selected by a crystal filter as for the other supplies. Thus 306 kilocycles are derived from 316 kilocycles (multiple of 4 kilocycles) and 10 kilocycles (multiple of 5 kilocycles).

The frequencies for the channel carrier supplies are developed in one bay capable of feeding 12 systems, and the group and pilot frequencies are produced in an associated bay capable of feeding 8 systems.

Units of the carrier supply system, which on failure would cause all 12 channels to fail, that is, the master oscillator, harmonic generator, and group carrier amplifiers, are duplicated; in the event of failure automatic change-over takes place. A further feature of this circuit is its ability to discriminate between a major and minor fault, for, whereas a failure of the 340-kilocycle supply would cause all the associated systems to fail, failure of any of the other group frequencies would cause only some of the systems to fail. In the event of a failure on both sets of supplies, the change-over panel functions to maintain working the set which has the less-important fault.

The repeater, shown schematically in Fig. 4, functions in the ordinary way to amplify the signals at the line frequencies and retransmit them to line at the appropriate level (+17 decibels relative to the transmitting trunk switch-board).

The method of regulation using two pilot frequencies in each direction of transmission is of interest. With changes in weather, especially in areas where fog, sleet, and ice are experienced, the attenuation characteristics of open-wire lines change in slope as well as in basic loss.

The gain-frequency characteristic of the repeater or receive terminal is designed to cover the ranges of slope and flat loss in the lines. As shown in Fig. 6, the slope can be varied between the limits defined by *cc′* (or *hh′*) and *dd′* (or *gg′*). Any slope characteristic inside these limits can be moved up or down within the range shown by changing the flat gain. It will be observed that the slope characteristics for any fixed flat gain come virtually to focal points at the highest line frequency in the low-frequency direction of transmission and at the lowest line frequency in the high direction. Thus, by using pilots at or near these focal points, the flat gain may be controlled, whereas pilots at or near the other extremes of the frequency ranges may be used to vary the slope without radically affecting the flat gain.

The pilots transmitted from the terminal are selected by narrow-band crystal filters at the output of the line amplifiers, at repeaters, or at the output of the receiving group amplifiers at receiving terminals. They are separately amplified and rectified, the direct component of voltage being used to bias an oscillator. This oscillator supplies power to the heater of a thermistor, which is a device producing a variation in resistance for a change in heater currents.

By incorporating the resistance element of the thermistor in the feedback or coupling path of an amplifier, the flat gain is readily controlled in the orthodox way. The variable-slope networks are designed to give a family of linear curves for different values of thermistor resistance. Thus, it can be seen that by using the appropriate pilots to control the slope and flat gain, it is possible to achieve the required characteristics shown in Fig. 6.

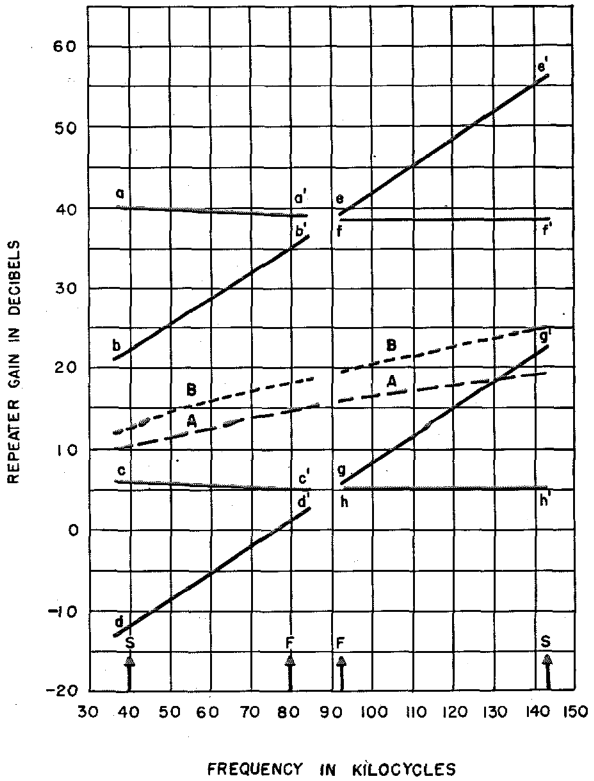


Fig. 6—Range of gain in repeater or receive terminal. *A* is the typical overall gain characteristic for 65 miles of 300-pound copper conductors spaced 8 inches in dry weather. *B* is for wet weather. *F* is for flat and *S* for slope pilots.

It will be apparent that the gain and slope variation in the transmission path are dependent on a change in pilot levels at the output. By suitably designing the circuits, it is practicable to cover the whole range shown in Fig. 6 by a pilot change of the order of 0.5 decibel. In other words, as the pilot-to-signal-level ratio is fixed, the signal levels will not vary by more than 0.5 decibel for all changes in slope and loss in the lines that can be covered by the range of the repeater or receiving terminal gain shown in the figure. For testing purposes, of if otherwise required, the regulation may be fixed manually by rheostats controlling the flat and slope gains of the circuits.

1.1 Alarm Trunk Circuits

On a system of this nature, with fully automatic pilots compensating for all reasonable line variations, it is possible to leave some of the repeater stations unattended. This is often desirable in view of the fact that repeater spacing of about 75 miles, which is normally practicable on the *SOJ* system, approximately doubles the number of stations previously used for three-channel operation.

The alarm trunk circuit has been developed specially for use in conjunction with unattended *SOJ* repeater stations. Faults or alarm conditions at the unattended station are indicated fully back to the attended control station, which may be the next repeater or terminal, by means of signals transmitted over a physical or derived direct-current telegraph circuit. Typical alarms indicated in this way include such conditions as mains failure and mains restored, rectifier failure, and low fuel in the engine-generator.

In addition, the unit operates in the other direction, giving the attended station control over the other station, enabling the remote attendant to switch power to a spare repeater, to switch from automatic working with pilots to manual operation of the repeater, or to effect other similar changes as required.

1.2 Interstation Communication

Facilities are provided whereby all stations or selected groups of stations along a route may be in direct communication. By using bridging filters across a suitably chosen line, the circuit is arranged to introduce no appreciable loss except when actually in the talking condition, so that there is negligible interference with normal commerical traffic. The effect at all times on a three-channel carrier system operating on the pair is negligible.

Any physical or derived direct-current telegraph leg may be used in conjunction with the telephone panel to provide a signalling circuit between the stations and any simple code may be used to make the calling selective.

2. *Line Facilities in the Union*

The administration has done a considerable amount of experimental work with a view to determining the type of line construction most suited to its needs. New main trunk routes are being built entirely to the *J*-3 transposition design, in which every pole on 6.4-mile transposition sections is a potential transposition pole, 300-pound copper wires are supported on 40 poles per mile, the wires of a pair being spaced 8 inches apart, and the crossarms being 24 inches apart. The pole spacing and sag regulation is kept within close limits as required by the theoretical design. On existing *K*-8 and *K*-8-2 routes, with 32 or 40 poles per mile, where *SOJ* operation is required new *J*-3 pairs are being erected, the existing wires being retransposed where necessary to avoid conflict with the *J*-3 transpositions. A typical terminal pole-head with lead-in cable for two *SOJ* pairs is shown in Fig. 7.

Measurements on the lines are facilitated by the use of special mobile test vans equipped with oscillators, transmission-measuring sets, crosstalk sets, impedance bridges, etc., together with the necessary power supply and lighting arrangements. Up to four open-wire pairs at a time are connected from the crossarms to separate insulated terminals on the exterior of the van by loosely twisted pairs of 7/0.029 indiarubber vulcanized wire, the terminals on the inside of the van being wired to a U-link panel. The test gear is wired to other U-links and switches, and the apparatus appropriate to the required measurement can thus be readily connected to the pair or pairs under test. Adequate accommodation is provided in the van for two observers.

In a country that comprises so many diverse geographical features, maintenance considerations dictate that the main routes should be readily accessible to wheeled transport, and they are, therefore, constructed along the national highways.

In the smaller towns, it is possible to run the open-wire route very close to the office. Specially developed low-capacitance star quad cable is used to lead in from two open-wire pairs down the pole and into the office. The cable is supplied with terminations for mounting on the line-filter bays in the office and on the crossarm of the pole. Short stubs of cable are sealed into the terminations, and the appropriate length of cable is jointed into the stubs at each end. Connections from the pole-mounted termination to the open-wire lines are made with bridle wire via the loading units and coil-and-protector boxes described later. This arrangement makes it convenient to break into the circuit and carry out tests and localize faults.

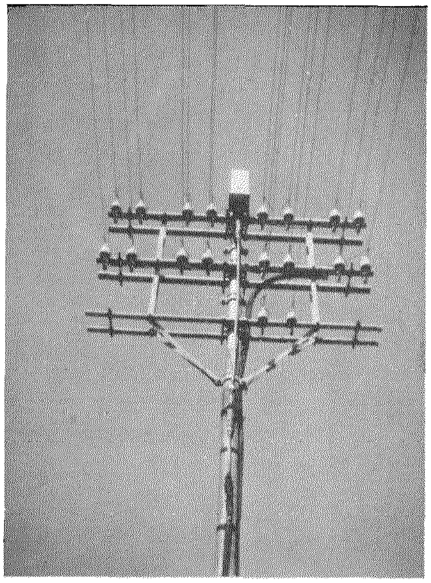


Fig. 7—Typical *J*-3 terminal pole with lead-in cable for two *SOJ* pairs.

To prevent reflected near-end crosstalk on the open-wire lines from becoming a controlling factor at the far end, it is necessary to reduce impedance irregularities to a minimum. For this purpose the lead-in cable is loaded so that its impedance closely matches that of the open-wire line. For lengths of lead-in up to 180 feet, the required degree of matching is achieved by suitably adjusting a variable load unit mounted on a line-filter bay in the office. For longer lengths up to 330 feet, this load is supplemented by a fixed load unit mounted on the crossarm of the terminal pole, connection to this unit being made by the bridle wires from the cable termination on the office side and by similar wires through the coil and protector boxes on the open-wire side.

The degree of matching achieved is illustrated in Fig. 8, which shows the reflection coefficient between the open-wire line and the lead-in cable at Johannesburg. The office end of the long entrance cable was terminated in its nominal impedance of 125 ohms for the purpose of these measurements. Normally, of course, with the equipment connected, the impedance of the directional filter would cause mismatch in the frequency range from 84 to 92 kilocycles.

In the larger centres, the open-wire route terminates some distance from the office, and the treatment is then necessarily different from that described. One method is to use the low-capacitance star quad cable with full loading at 600-foot intervals, the terminal loads being designed for mounting on the pole crossarm at one end and on the line filter bay at the other.

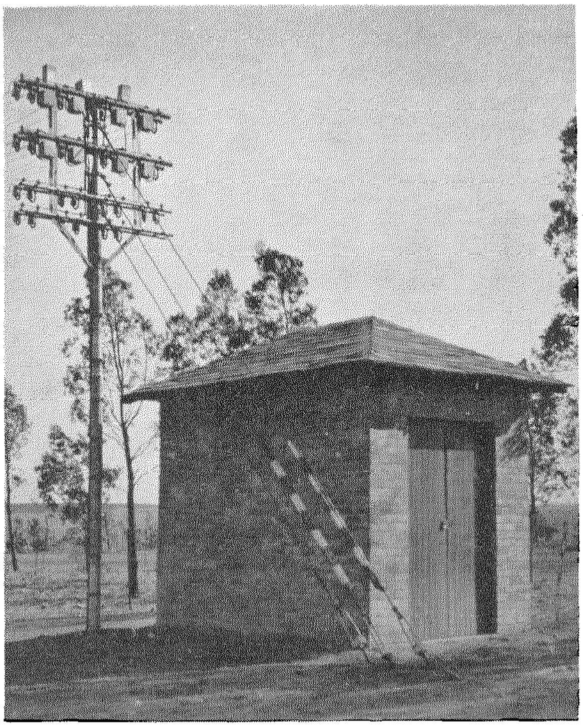


Fig. 9—Filter hut at Kroonstad.

The administration, however, has standardised for economic reasons on an alternative method using filter huts. It is not economically practical to develop a satisfactory loading system for paper-insulated toll entrance cables of the usual capacitance at *SOJ* frequencies. If unloaded pairs are to be used, it becomes necessary to match the impedance of the cable to that of the open wire and equipment. As the impedance of such cable changes markedly with frequency below about 20 kilocycles, it is not practicable to produce a suitable matching transformer for high-grade operation at all frequencies in the voice and carrier range. Such a transformer, however, is available to cover the *SOJ* range of frequen-

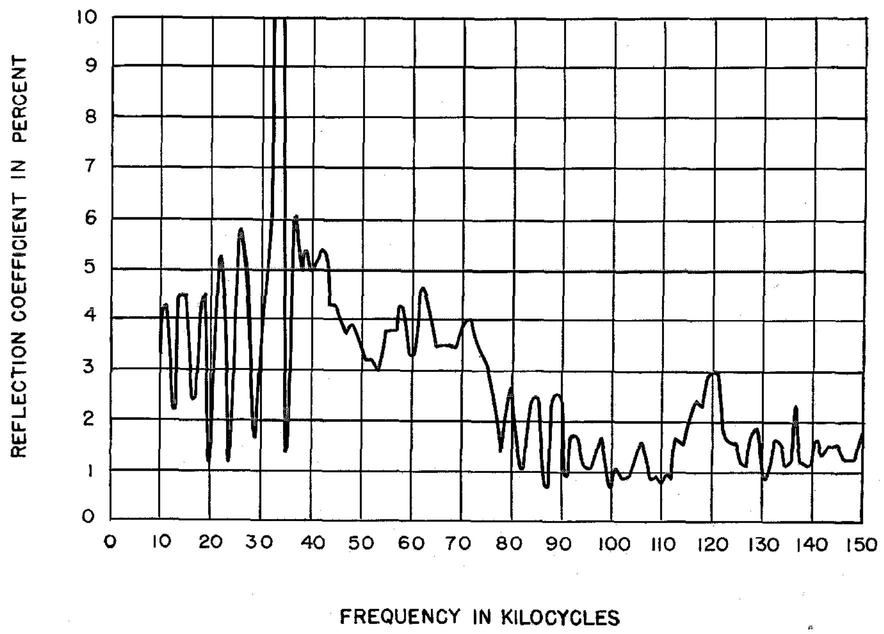


Fig. 8—Reflection at junction of open-wire line and lead-in cable at the Johannesburg terminal.

cies, and the procedure, therefore, is to separate the *SOJ* from the lower-frequency systems on the circuit, leading the former into the office over a non-loaded pair with impedance matching coils and the latter, as hitherto, over a pair loaded for three-channel operation. This separation is achieved by housing the normal *SOJ* line filters in a filter hut erected near the junction of the entrance cable and the open-wire routes. The existing entrance cable at Johannesburg is some 8000 yards long, and extensive tests carried out by the administration indicate that 10,000 yards is a reasonable practical limit. Over this length in a 54-pair multiple twin cable, it should be possible to select 8 pairs that can be balanced for satisfactory crosstalk. The lead-in arrangement on the *SOJ* pairs from the pole crossarm to the hut are precisely the same as those already discussed. Fig. 9 shows the filter hut and terminal pole on the north side of Kroonstad.

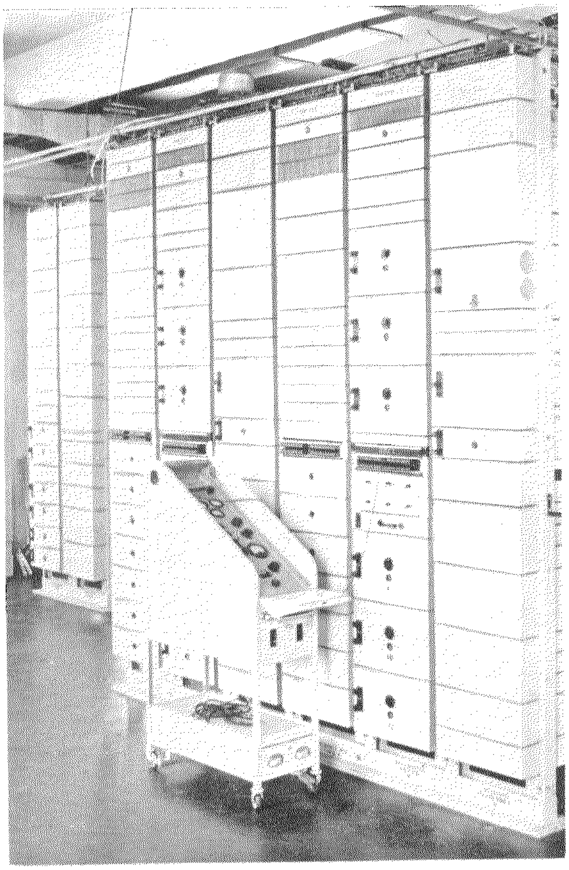


Fig. 10—Part of the Bloemfontein terminal equipment.

To limit crosstalk from the *SOJ* into other pairs or into the longitudinal path, retard coils are inserted in the wires. These coils, together with protectors, are mounted in a box on the terminal-pole crossarms, connections to the open wires and lead-in cable (if necessary via the load) being made with bridle wires as mentioned previously.

As severe lightning storms are frequent in the Union, the ordinary protection is supplemented by drainage. The arrangement comprises a coil bridged across the line in series with protection on each side of the coil. The coil centre point is earthed, and the protectors operate at a lower voltage than the main-line protectors. Hits that would not cause the normal protectors to break down can seriously affect the operation of voice-frequency telegraphs, and the purpose of drainage is to reduce these disturbances by ensuring that the two pairs of protectors in series with the coil break down together and thus minimise the currents in the metallic circuit. A subsidiary function is to prevent intermittent breakdown of the main-line protectors during severe dust static. The apparatus is mounted in a box identical with that used for the longitudinal coils and protectors.

3. *Equipment*

The equipment follows the standard practices, being mounted on both sides of rack frameworks 10 feet, 6 inches high, on panels 19 inches wide. Fig. 10 shows some of the equipment at the Bloemfontein terminal. The two carrier supply bays can be seen on the left, the others from left to right being ringer, channel and frequency-translating bays for system 2, and duplicates for system 1.

Fig. 11 is a photograph of the equipment in the unattended repeater station at Winburg. Working and spare repeater bays are shown in the left foreground with one line-filter bay immediately behind and the edge of the other on the right. In the background, adjacent to the toll test board, is the miscellaneous-apparatus bay mounting the alarm trunk unit for a controlled station.

The introduction of these systems in the Union has, in most instances, entailed the erection of

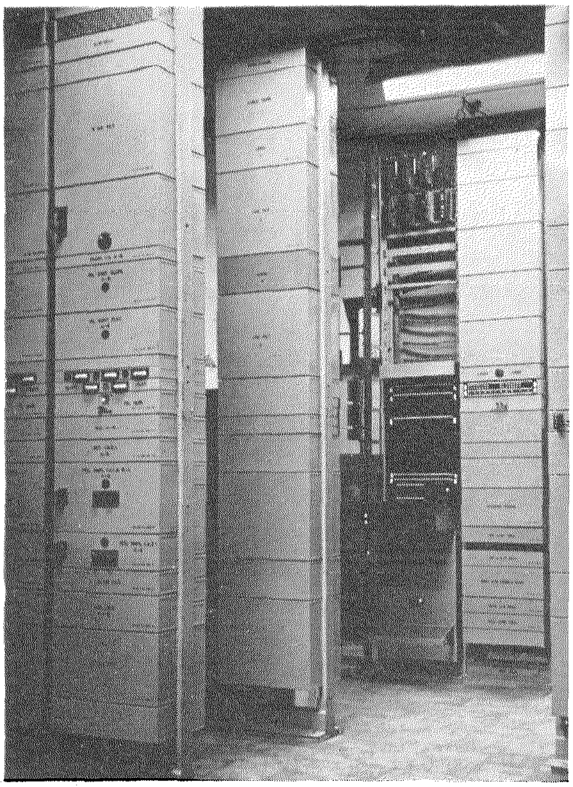


Fig. 11—Repeater equipment at the unattended station at Winburg.

entirely new buildings, either on account of the reduced repeater spacing necessitated by the introduction of *SOJ*, or because the existing carrier accommodation has become inadequate to deal with the new and expected developments. With one or two exceptions, the repeater buildings follow a standard pattern, the plan of which is shown in Fig. 12. It can be seen that the accommodation has been planned on a very liberal basis to allow for considerable development. A typical photograph of such a station is shown in Fig. 13.

The accommodation required in large terminal stations cannot be standardised to the same extent, and each one has, therefore, been considered and planned individually, but always with a view to future development.

All the filter huts are new buildings, and the administration has standardised on two sizes, the larger of which will cater for 16 pairs and the smaller for 8 pairs. The bays supplied for installation in the huts are 7 feet high. The first bay mounts terminations for 8 *SOJ* pairs (4 quad cables) and line filters and loads for 2 pairs. The second and third bays each mount 3 line filters and loads, so that the 3 bays cater for 8 pairs.

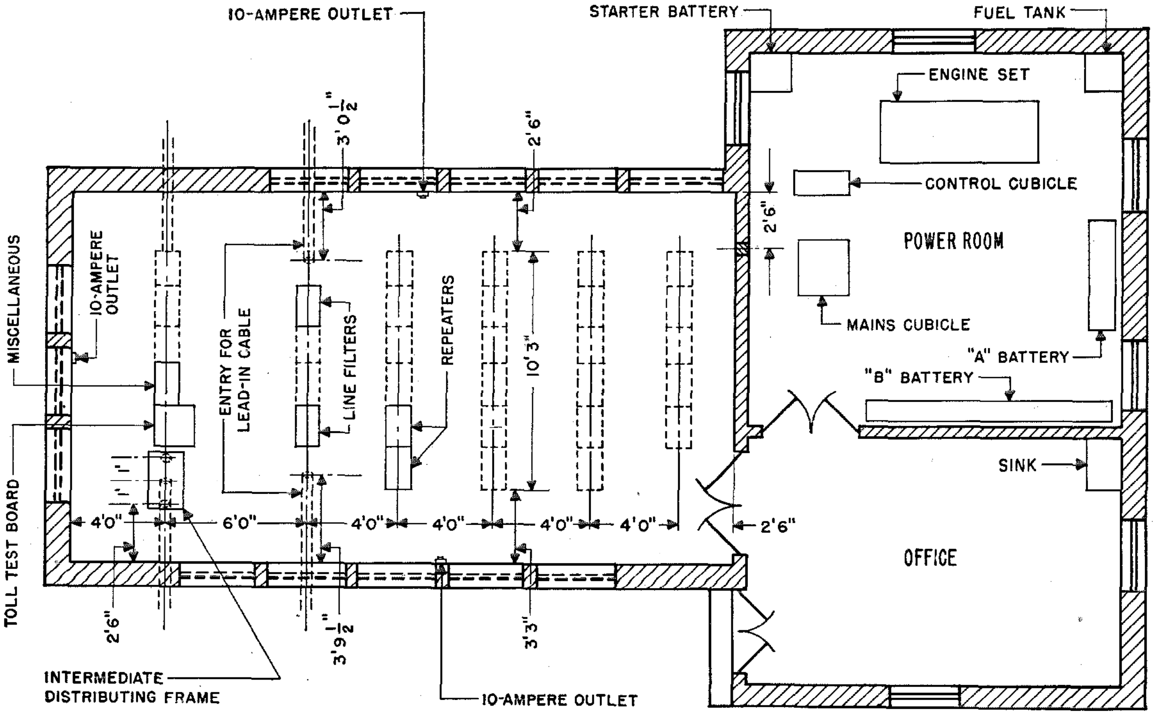


Fig. 12—Standard repeater station arrangement.

3.1 TEST GEAR

The test gear supplied with the equipment for maintenance purposes is described briefly in the following paragraphs.

At terminal stations, a trolley carries most of the facilities normally required for routine maintenance. The form of this item can be seen in the foreground in Fig. 10. It comprises the following apparatus mounted on a steel framework.

A. Transmission-measuring set with all necessary sending and receiving facilities over the *SOJ* frequency range.

B. 800-cycle oscillator.

C. A 14-fixed-frequency oscillator giving 12 frequencies equivalent to voice frequency on each channel at the output of the channel modulating equipment and 2 frequencies corresponding to middle frequencies in the bands transmitted to line in the two directions of transmission.

D. Meters for voltage and current measurements and a U-link panel.

The trolley operates from the normal equipment supply voltages, 24 and 130 volts, and each bay for the *SOJ* equipment has a socket suitable for the connecting plug on the trolley.

For measuring the levels of the group-frequency supplies, i.e., up to 600 kilocycles, a small high-frequency voltmeter is provided with 0–2.5- and 0–5-volt scales. This set is all that is required at these frequencies for routine maintenance measurements. For more detailed tests in cases of breakdown, however, a mains-operated transmission-measuring set, reading down to −55 decibels in the frequency range from 60 to 600 kilocycles is supplied.

At repeater stations, a portable transmission-measuring set, which obtains its power supplies via a cord which can be plugged into the sockets on the bays in a similar way to the trolley, is supplied for maintenance purposes. This transmission-measuring set incorporates all the necessary facilities on the send and receive units for such measurements and an oscillator giving the two frequencies corresponding to middle frequencies in the bands transmitted to line in the two directions of transmission. The unit can be used with 12- and 130-volt batteries, when required for measurements in the filter hut or on the lines.

Apart from the gear described above for routine and maintenance tests, the administration has felt the need for apparatus suitable for making more precise measurements on detailed parts of the equipment. For this purpose at terminal stations of combined 3-channel and 12-channel repeaters, the following items are supplied:

A. Continuously variable heterodyne oscillator with a maximum output of +30 decibels, referred to 1 milliwatt, giving a high-grade performance at frequencies in the range 500 cycles to 150 kilocycles.

B. Transmission-measuring set covering the frequency range 100 cycles to 150 kilocycles reading from −45 to +35 decibels, referred to 1 milliwatt, on the receive unit and incorporating a send unit suitable for use in conjunction with item A.

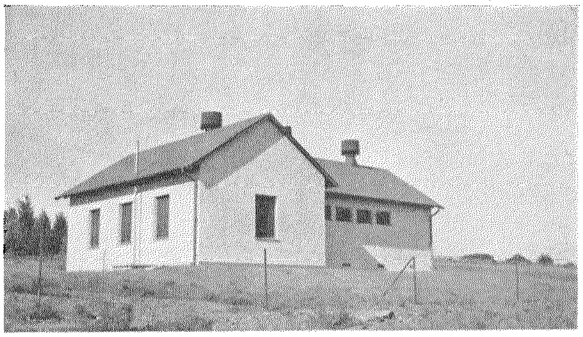


Fig. 13—Typical repeater station building.

Both units are operated from the alternating-current mains and they are mounted, together with the mains units, on a trolley framework.

4. *Power Plant*

As many of the stations on the *SOJ* routes are entirely new, no power equipment exists. In many of the others, the present power plants are inadequate to deal with the additional loads. The type of power plant used exclusively by the administration in these cases at repeater stations and the smaller terminals will be described. At the large terminal stations, generators with floating batteries are used.

The equipment supplies of 24 and 130 volts are normally provided by rectifier units operating from alternating-current mains. The units are designed so that variations of up to ±10 per cent in the mains-supply voltage or considerable variations in the load on the 130-volt supply are automatically compensated to maintain the output voltage constant within close limits. This

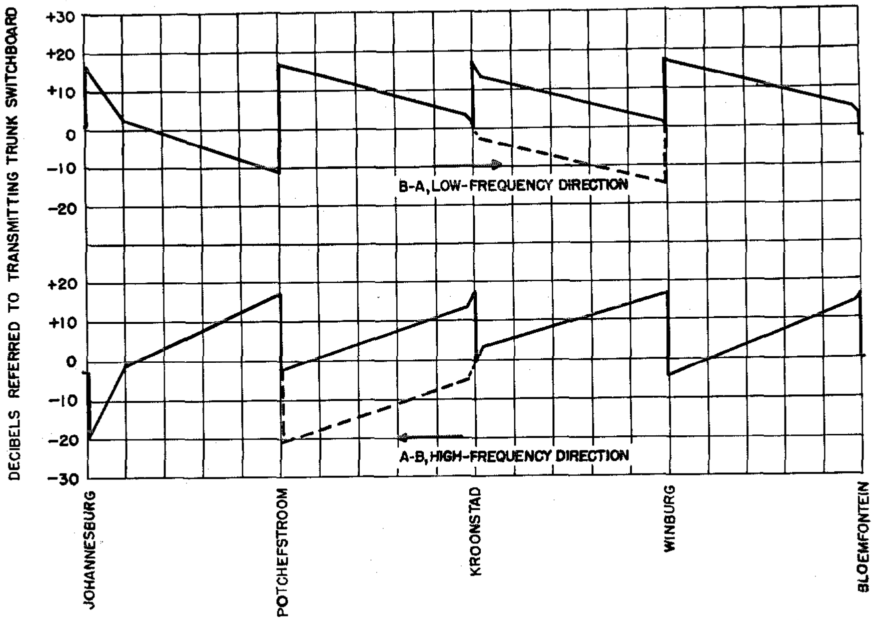


Fig. 15—Signal levels at the highest line frequency used between Johannesburg and Bloemfontein. The dotted curve is obtained with the lines patched through at Kroonstad.

is achieved by a motor-driven tap on the mains input transformer, the motor being actuated by a robust, sensitive, marginal relay in the 130-volt circuit. The transformer is designed so that any correction necessary on the primary side to maintain the higher voltage within the required limits due to changes in mains-supply voltage also causes a corresponding change in the lower voltage, thus maintaining this within approximately the same close limits.

Similarly, if the higher- and lower-voltage loads vary in the same ratio, the lower voltage is maintained approximately correct. As the *SOJ* equipment is designed to use ballast lamps in all the valve heater circuits, considerable divergence in the load ratios is permissible before the lower voltage falls outside the range that can be compensated. The higher voltage will be maintained within the limits of the control relay and, in practice, it is possible to achieve approximately ±1 per cent.

Three sizes of rectifiers are available, delivering at lower and higher voltages, respectively, 30 and 2 amperes, 60 and 4 amperes, or 100 and 6 amperes. The higher- and lower-voltage supplies in the two smaller plants are derived in one cabinet, but for the larger plants, separate units are used. A photograph of a 60/4 unit is shown in Fig. 14. The other equipments are similar in appearance, the height and width being the same in all cases.

As protection against mains or rectifier failures, the standby plant consists of a Diesel engine-generator yielding regulated direct-current outputs at 24 volts and 130 volts to the station

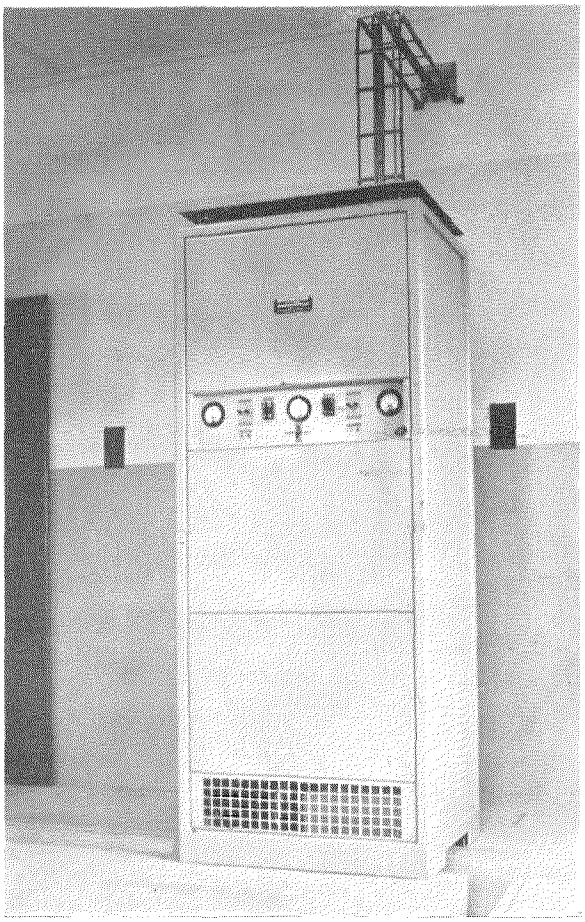


Fig. 14—*Rectifier delivering 60 amperes at low tension* and 4 amperes at high tension.

equipment through the smoothing filters in the rectifier. This plant is arranged to start up automatically in case of a mains or rectifier failure. Small-capacity batteries of a totally enclosed type are floated across the rectifier outputs and supply the equipment during the period of some 15 seconds before the generator takes over the load. The engine starter battery is fully recharged from the starter motor, which becomes a generator following the automatic change-back of the load to the rectifier unit on restoration of the normal mains supply, before the engine shuts down. A fuel tank, capable of keeping the engine running for 48 hours, ensures that even at unattended stations there will be no break in the service. It is considered practicable within this time for an attendant to reach the station and either refuel the engine or service the power plant. With such a reserve, this action should be possible even at week-ends and similar periods, when only a skeleton staff normally mans the station.

5. Line-up of Circuits: Joannesburg-Bloemfontein

Fig. 15 shows the signal levels, relative to the transmitting trunk switch-board, at the highest line frequency along the route in the two directions of transmission, as measured in dry weather during the initial line-up of the system. The greater slopes at Johannesburg, Bloemfontein, and on both sides of Kroonstad are due to entrance cables, but it should be understood that no attempt is made to show the relative lengths of cable and open wire on the horizontal axis. The dotted lines show the levels with the lines patched through at Kroonstad; during the course of the tests such a patch was made and the system was found to align itself and work perfectly satisfactorily. By examining this figure in conjunction with Fig. 6, it will be seen that even under these conditions the repeater still has a considerable margin of gain in the two directions.

The circuits were lined up to a 3-decibel 2-wire equivalent. It is interesting to note that ultimately it will be lined up to 0 decibel, for, to conform with the administration's zoning plan whereby any two subscribers throughout the country may be interconnected with a prescribed maximum overall loss, all major trunk carrier circuits will have a zero equivalent. Automatic trunk switch-boards providing 4-wire switching on a tail-eating basis with pad connections will be used.

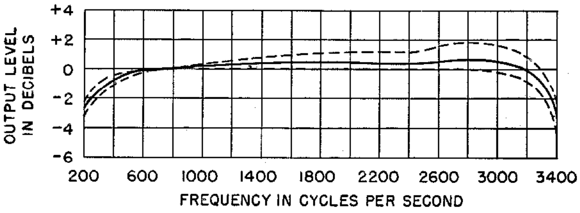


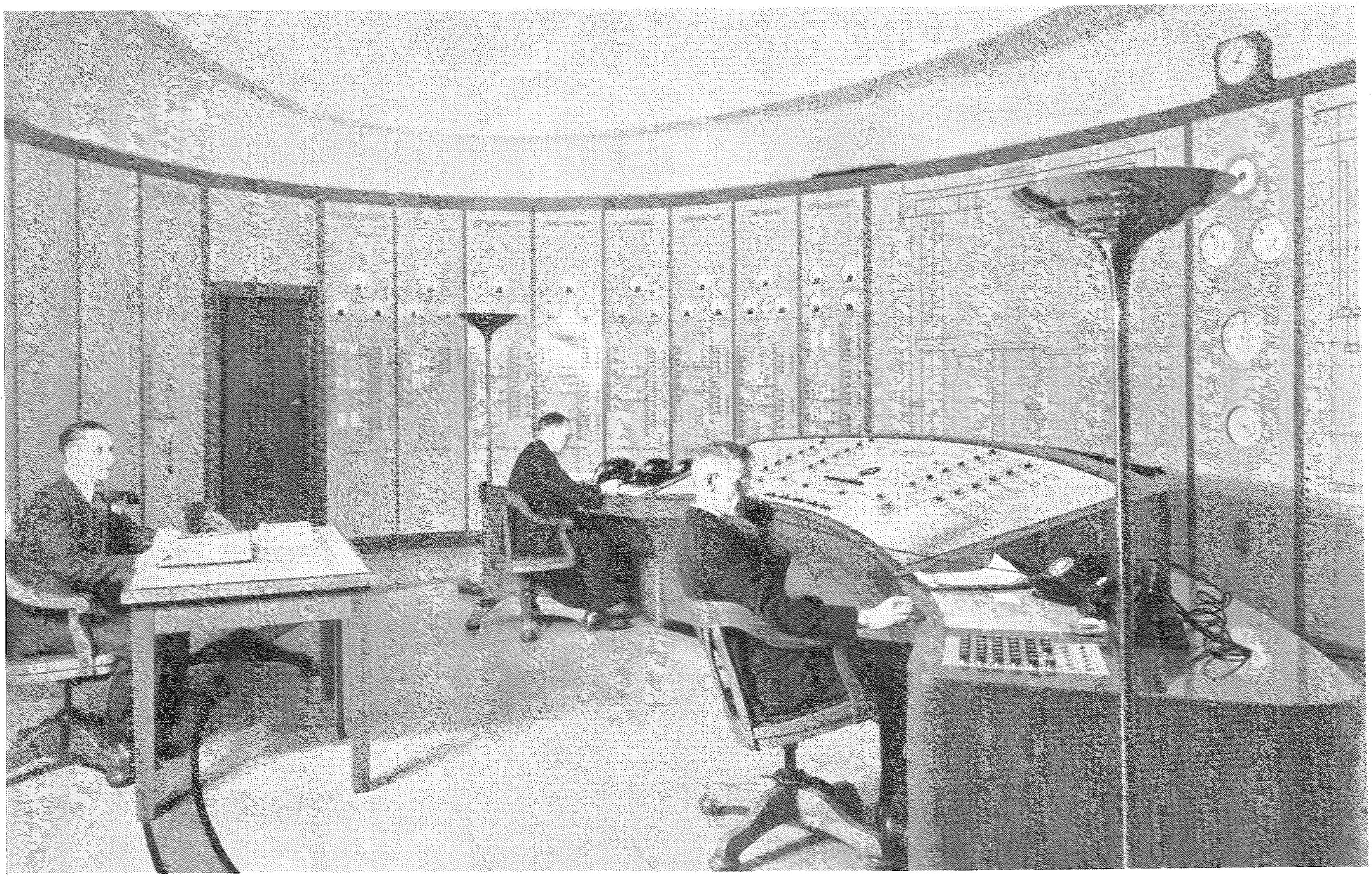
Fig. 16—Quality curve related to 800-cycle transmission. The solid curve is the average of 12 channels and the maximum departures are indicated by the dotted curves.

Fig. 16 shows the quality curve, relative to the 800-cycle point representing the average of the 12 channels in the B-A direction of transmission. The dotted curves show the spread, i.e., they are drawn through the extreme figures at each frequency.

Noise measurements were made on each channel, after lining up to 3-decibel equivalent in each direction of transmission. The worst channel gave 2.0 millivolts psophometric electromotive force of weighted noise. All other channels gave figures better than 1.0 millivolt, more than half the channels being better than 0.5 millivolt.

6. Acknowledgment

The authors express their appreciation to all those who, directly or indirectly, made possible the writing of this article.



General view of the central control room of the Manchester Corporation Electricity Department. In the background are the panels for the remote control, indication, and metering of nine 33-kilovolt main substations. To the right of these is the 33-kilovolt system diagram separated from the 6.6-kilovolt diagram, shown in part on the extreme right, by a central loading and frequency panel.

Operational Control of Electricity Supply Systems*

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SUPERVISORY EQUIPMENT for the remote control of plant has proved to be thoroughly reliable and to facilitate efficient operation of electricity supply systems. This paper gives the reasons for, and the steps taken to develop, the common-diagram control system, which enables an almost unlimited number of substations, etc., to be completely controlled from one diagram and control panel and is sufficiently flexible to cater for the growth of the undertaking. It describes a wall-type system diagram which automatically indicates which substations have changed conditions, and therefore the area involved in any disturbance. The system diagram is equally extensible to accommodate new feeders and substations, with a minimum of operating disturbances.

Particulars of the circuits and apparatus, and comparisons of floor area, pilot, and cost economies are given; also information of an installation dealing initially with 78 substations, to which others are being added.

• • •

1. Introduction

1.1 GENERAL

Factors essential for the satisfactory operational control of an electricity supply system include:

A. Accurate forecasts of load conditions.

B. Knowledge of current flow and fault MVA in all parts of the system.

C. Immediate notification of occurrences on the system.

D. Facilities for voltage regulation and the performance, without delay, of switching operations.

Many supply authorities have an organization for the collection of statistics, and for the planning and execution of operations. Calculating boards to ascertain fault MVA and the current flow, in all parts of the system under all conditions of working, are of great value in enabling the results of switch operation to be forecast. The more progressive organizations have installed remote supervisory control gear for the rapid acquisition of information from the system, to enable voltage to be regulated and switching performed directly from a central control room.

1.2 REASONS FOR DEVELOPING A COMMON-DIAGRAM SUPERVISORY SYSTEM

Supervisory control equipment which was in service prior to the year 1943 had the disadvantage of being found inflexible when extensions were necessary. If a larger number of substations had to be controlled, the floor space required became excessive because an individual panel was required for each substation, whereas for the common-diagram system only one small desk is required for any number of substations, and only a pair of pilot lines is required for a group of substations (Fig. 1).

It became obvious that, to render practicable the application of remote-control equipment to a large number of substations, new development beyond these individual systems was essential.

Routine operational control is secondary in importance to that which it is necessary to provide for a network disturbance. For this condition a control engineer requires to know first the extent of the disturbance, and second, detailed information of the state of affairs at various points on the network. This detailed information he will acquire point by point in order of importance as judged from his experience and knowledge of the extent of the disturbance.

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Therefore, if there is a pictorial representation of the whole system which indicates automatically the substation affected by a disturbance, then the engineers' requirements will be fulfilled by a "common" control desk providing detailed control and indication of any substation.

This paper describes a new supervisory-control system designed to meet these requirements for Manchester. It includes a wall-type system diagram in combination with a control desk of moderate size, which is capable of having any additional number of substations connected to it, without increasing the size of the desk, and at less cost than individual panels.

1.3 System Diagrams and Control Boards

A large diagram with miniature indications automatically displaying every switch position in the system might seem to be ideal, but a few observations on its undesirability will illustrate the reasons for choosing the type of diagram adopted.

If switch indication only is required, the diagram may be 10 feet high and the indicators will be within vision, but if control is also required, experience has shown that the control switches should be located between 2 feet and 6 feet 6 inches above the floor level. This consideration means that, if many feeders and substations are to be represented, the diagram becomes disproportionately long, to the detriment of its pictorial value, and, owing to crossing lines, results in loss of clarity.

Individual automatic switch indicators and control keys require individual relays and wiring. This tends towards inflexibility, and, although mosaic types of diagrams are helpful in overcoming this defect, it is doubtful whether they could successfully cater for both control and indication. Diagrams should therefore be capable of rapid and easy modification, but this becomes increasingly difficult of attainment as the amount of detail on the diagram is increased.

A control engineer is less concerned with the geographical position of a substation than with its electrical position, i.e., the section of generators from which it is normally fed, the alternative means of supply, which feeders may be paralleled and which should not be coupled, etc.

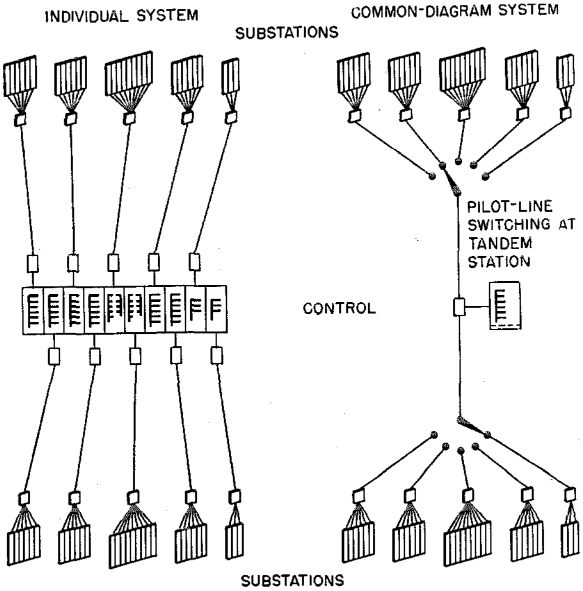


Fig. 1—A comparison in outline of individual and common-diagram systems. Supervisory equipment, indicated by the open boxes, is connected to the switchgear, illustrated by the larger boxes which are subdivided by vertical lines.

Even large, complicated systems can be represented by a diagram restricted to straight lines interconnecting substation symbols, each of which has a lamp automatically illuminated when the conditions at that substation change. These diagrams, built up of small moulded panels, can be arranged to form control room walls. They are built up on a jack-in tile basis, each tile being easily removable. Adhesive coloured tape is used to make rectangles and straight lines to represent substations and feeders, respectively. The tape diagram is easily and quickly modified to cater for extensions to the system. A pilot lamp is inset in the portion of a tile within the substation symbol. Connections to three such lamps can be made through four pin connectors per tile, which also form the means of fixing the tile. The system diagram has small plug-in disc indicators of various colours and symbols. This enables the wall diagram to be hand-dressed to represent, to any desired extent, the conditions on the system.

The geographical layout of feeders and substations forming a section of the area is shown in Fig. 2, and the form in which this section appears on the system diagram is shown in Fig. 3.

1.4 Common-Diagram Control Desk

More detailed information from a substation, and also remote control of equipment, must be provided if the system is to be complete. The control and indicating equipment must be capable of being applied at will to any substation.

When the substation plant arrangements for a distribution system are reviewed, it is found that the least common multiple of equipment required is not much larger than that of the ultimate largest requirement of any substation to be provided on the system. This is reasonably easy to state since, for reasons of economy, and also of load distribution and voltage drop, it is not good policy to permit the unlimited growth of a single substation, however great the load density may be.

Thus, a least common multiple of supervisory remote control equipment can be designed for any group of substations. The problem of making such an equipment available to any one of a number of substations involves two problems, namely:

A. Switching pairs of pilot wires to make them available to the equipment, one pair at a time.

B. Setting the equipment so that for a period of time it loses its least-common-multiple character and takes on that of the particular substation.

A solution has been achieved by the use of telephone-switching apparatus and methods, and may be summarized thus:

A. The diagram must represent the least common multiple of the substations, either by rows of lamps or by diagram.

B. Feeders and transformers, etc., must bear identity numbers, and means to display them.

C. For flexibility in extension, substations must be given numbers.

D. Means should be provided to display identity numbers of more than the one substation which is occupying the common diagram. In addition to lamp indication on the system diagram of all substations reporting trouble, there should

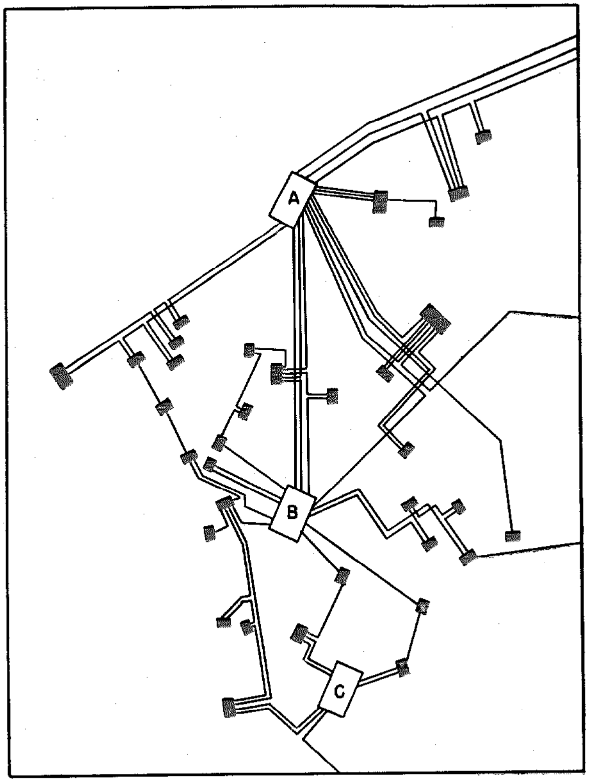


Fig. 2—Geographical layout of a section of feeders and substations.

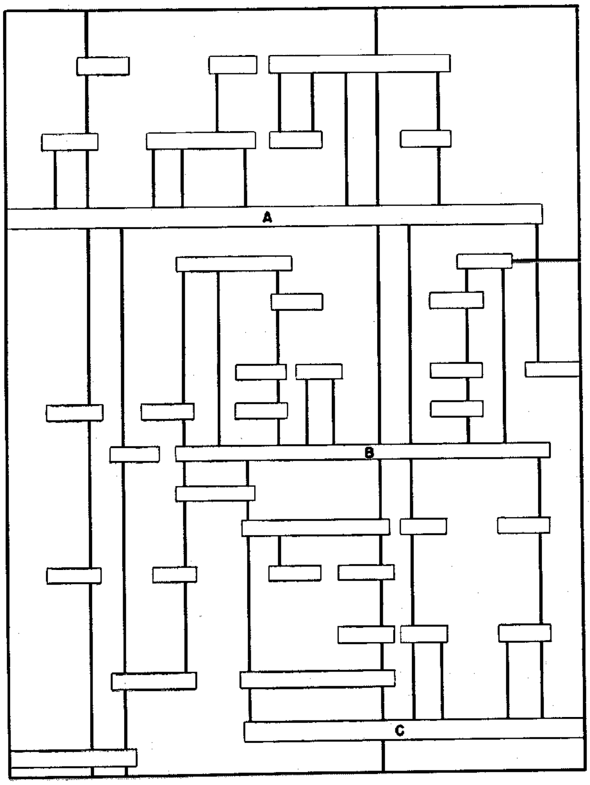


Fig. 3—System diagram layout for the section of feeders and substations shown in Fig. 2.

be means to display the identity number of substations occupying or requiring to occupy the common diagram.

1.5 Pilot Lines

Line-signalling circuits between the control station and the substations are necessary in any supervisory system. The pilot lines which exist between substations are rarely the result of a single planned installation which considers all possible present and future requirements of telephony, protection, intertripping, remote control, etc. The two extreme arrangements possible in any set of pilot lines are:

A. Pilot lines radiating individually to all substations from the control point.

B. All substations connected on one omnibus pilot line to the control station.

Rarely would either be met on a large system. This results in each case having to be treated on its merits, but a study of almost any existing network usually reveals a possible satisfactory arrangement even if, in a few instances, lines have to be added or switched. In the common-diagram system, economy can be effected by grouping substations around radiation points so that only a one-pair line between each substation radiation point, called a tandem station, and the central control is required.

It might be argued that it provides bottlenecks if a number of substations require to signal a supply failure simultaneously, but this is not serious for two reasons:

A. On any scheme employing a common diagram, fundamentally only one station can be connected at a time and therefore a pilot-line bottleneck at each tandem station does not worsen this arrangement.

B. Also, substations can be arranged to signal their identity immediately a supply failure occurs. This signal is short, and when received it lights the station pilot lamp on the system diagram. Immediately one such signal is received, the equipment is free to accept a second, so that as little time as possible is lost in providing the control engineer with the names of the substations affected by the failure. He, then, from his knowledge of the network, assisted by the system diagram, should be able to determine in what order of priority he desires to examine the individual switchgear conditions of the substations and select and call them to his common diagram in that order.

Indeed, this method ensures that the control engineer operates in the correct manner in the event of a major disturbance, in that it first directs his attention to the overall picture before it is too strongly attracted by one particular feature about which he might be tempted to take some premature action.

Further economy could be effected by the adoption of party-line working from each tandem, but as the number of substations for each tandem may be larger than present-day relay design allows for in party-line working, a mixed system from each tandem would result. It is better for circuit uniformity to have either all non-party lines or all party lines; the layout of existing pilots will in general govern the decision. For example, common-diagram system applied to the control of traction substations would require party lines rather than the tandem-station arrangement to provide an economy of pilot lines, whilst supply distribution will in general find the tandem-station, non-party line, best suited to its geography.

The circuit design can, if desired, be arranged to permit the operation of a telephone over the same pilots as the supervisory facilities.

1.6 General Plan

Association of these ideas enables a general plan to be stated for their application to a specific case.

Each substation is connected by a telephone pair to a tandem station. Equipment at the tandem consists of line relays and a line-finder, so that the pilots connecting the tandem station to the central control equipment can be switched to any one of a number of substations at the request of either the substation or the control engineer. One telephone pair only is required between each tandem station and the control station.

Circuits at the control station are so arranged that each tandem-station line circuit can be connected to a free connecting or link circuit under stimulus from either a substation equipment or the control station. At the same time a substation marker circuit can be energized by information received over the tandem-station line circuit, indicating the number of the substation which is calling or is to be called.

By these means, a substation is identified and its number displayed on the desk either by the control engineer originating the request for it or by fault conditions causing the substation to indicate these conditions.

Simultaneously, the particular substation lamp on a system diagram is illuminated, the lamps showing the extent of the area affected by a disturbance. The number of lamps that can be illuminated simultaneously is independent of the number of links serving the desk, and, whether or not the desk is free, an incoming signal from a substation must always serve the purpose of illuminating its associated lamp on the wall-system diagram.

Having occupied a link circuit, a substation can be arranged to occupy the common diagram on the control desk, causing it to display the switchgear conditions at that substation, and to provide remote-control facilities.

2. *Application of the Common-Diagram System*

2.1 General

The application of common-diagram supervisory equipment to the operational control of an electricity supply system can now be considered, Manchester being taken as an example and the number of 33-kilovolt and 6.6-kilovolt substations being 14 and 410, respectively. The 33-kilovolt substations are used as control tandem stations.

Supervisory control of some of the non-attended 33-kilovolt substations had been in commission since the year 1929, and the experience of operational control proved that the equipment was a valuable asset; supply had been resumed within a few minutes after interruptions, whereas without supervisory gear a delay of half an hour would have occurred. It was decided to extend the use of supervisory gear; a selection of substations (78), where load and switch control was considered to be of particular importance, was made for an initial installation. A proviso was made that the type of gear to be installed must be capable of extension to deal with a much greater number of substations.

The individual-panel type of supervisory gear was, for reasons previously given in the paper, considered to be impracticable for such large-scale development, and consequently the common-diagram system was evolved.

2.2 Problem of Relating All Types of Substation to a Least Common Multiple

In most undertakings there exist variations in the design of substations, but these physical differences do not seriously affect supervisory control and indication.

Any system of remote switchgear indication is concerned primarily with three fundamental types of equipment:

A. Two-position devices such as oil circuit breakers, isolators, alarms, etc.

B. Multi-position devices such as transformer tap-change gear.

C. Meter readings or the transmission of continuously variable quantities.

The controls required are almost wholly confined to one of two conditions, namely "closed" or "open," "on" or "off," "raise" or "lower."

From these fundamentals many variations are derived, such as duplicate busbars, "fleeting" contact alarms, etc., all of which usually involve only the addition of relays to translate their variations back to the standard types of condition.

In designing a control desk, therefore, to cater for various types of equipment, the same control and indication devices can be used, since these have only to fulfil one or other of the above-named universal conditions. Similarly, so far as the control panel is concerned, the type of the switchgear and diagram of its operating circuits are immaterial; where differences occur they must be catered for by local supervisory relays.

The common-diagram control desk therefore includes the greatest number, in any one substation, of two-position devices that will occur, together with a method of indicating to which feeder, transformer alarm, etc., each device belongs; and similarly for multi-position devices and meters.

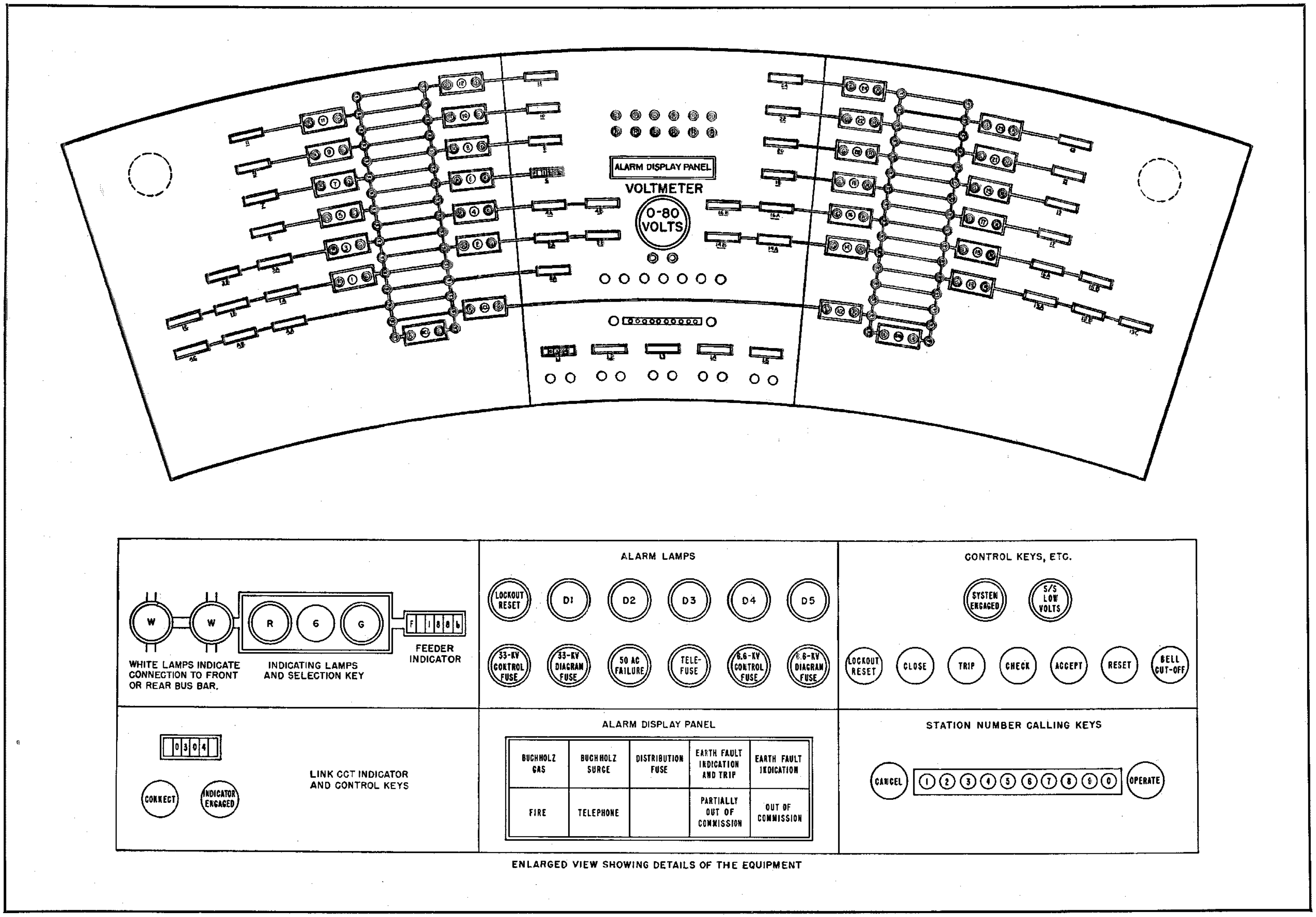


Fig. 4—Common-control-diagram layout.

2.2.1 Display of Switchgear Identity

The problem of indicating the name or number of the device beside the indicating lamps and control keys can be solved in several ways, but to be universally flexible and extensible the scheme should be confined to the use of letters and figures representing a simple code of definition, such as: T1 for Transformer No. 1, F1 for Feeder No. 1, etc. Such a code, to almost any variety, can easily be set up and displayed on digit-wheel indicators.

Thus, given a control key, indicating lamps, and electrically operated digit-wheel indicators forming a unit for a two-position control and indication, a number of them could be associated at will to represent any arrangement of switchgear.

2.2.2 Busbar Arrangements

Further, if the substations have either single or duplicate busbars, these indicating units can be mounted on the control desk one above the other and shown diagrammatically connected to the busbars. Since any unit can be designated by any indicator, the order of connection to the busbars can be suited to any substation.

Busbar couplers and section switches can be catered for in the diagrammatic representation by placing these units in position and working along the busbars on either side as far as it is necessary to go for any particular substation.

Many substations will not use all the units provided; this will not cause confusion, because the lamps of only those units actually in use at any instant will be illuminated. Additional busbar-selection indicating lamps, remotely operated, will show the connections of the switchgear to the busbars obtaining at any time.

2.2.3 Indication of Substation Numbers

To inform the control engineer which of a number of substations is using the diagram, the substations must be numbered and displayed on a stepping numerical indicator or similar device. Either an instrument similar to a telephone dial, or a row of digit keys, will enable one of an infinite number of substations to be selected.

2.2.4 The Common-Diagram Desk

The equipments in the Manchester substations at present vary from one feeder, one transformer, and one circuit breaker, to 12 feeders and 7 transformers on 11-kilovolt and 6.6-kilovolt single- and duplicate-busbar sectionalized boards. In some substations having two feeders and two transformers only, one circuit breaker is installed. The complete lowest common multiple of facilities provided is listed in Table II, whilst Fig. 4 shows the common diagram which was developed for indicating and controlling the combinations of equipment in the present and future substations.

Duplicate busbars on each side panel are provided with busbar couplers and joined by an interconnector towards the front of the panel. Beneath the layout of the panel, six enlarged views illustrate the items of equipment, including that for a two-position remote control and indication; the series of alarm lamps (positioned above the alarm display panel); the common control keys (positioned below the supervisory system voltmeter); details of the connecting links for the display of substation numbers (five of which were provided and positioned across the front of the centre panel); particulars of the other remote alarms provided (illustrated as they are displayed when illuminated in the alarm display panel); and the digit keys used for "dialling" the substation number, together with the "operate" and "cancel" keys which render the dialling effective or otherwise.

Since this common-diagram installation was put into commission, arrangements have been made for the indication and removal of "lock-out" for 26 trolleybus supply circuit breakers in 11 substations, whilst earth-fault alarm indications are being provided from an additional 37 substations.

2.3 PILOT-LINE ARRANGEMENTS

There were pilot-line pairs available in the telephone network between most substations in an area and the 33-kilovolt substation supplying that area; also a single pilot pair between each 33-kilovolt substation and central control. It was therefore decided to install tandem-station working, 12 circuits radiating from central control to the 12 tandem equipments (at the 33-kilovolt substations), which enabled these circuits to be extended to any one of a series of pilot pairs connecting to the substations in the area, either by signals from the substations or by a coded signal from the central control.

2.4 SUITABLE SUPERVISORY REMOTE-CONTROL SYSTEM

Methods of employing a common-diagram system depend upon a technique well known to automatic-telephone switching engineers, for, given the method of supervisory signalling, the rest is a matter of calling a particular substation, or being called by it, and connecting its line wires to the common-diagram equipment. It is obviously convenient to make the impulsing voltages used for the supervisory and substation calling signals the same, but, in general, almost any method of supervisory-control signalling could be employed which employs impulse trains for selection and does not depend upon maintaining a continuously rotating synchronism between substation and control.

To illustrate the method by which a large number of stations are remotely controlled and supervised from one control desk, a basic method of supervisory control will be explained and then the expansion of this method into a common-diagram system developed.

3. *Common-Diagram System*

3.1 BASIC SUPERVISORY REMOTE-CONTROL SYSTEM

In the various well-known methods of supervisory remote control which use impulse trains to perform the required selections, the detailed differences consist in the methods adopted to ensure that the impulsing is correctly transmitted and received, and in the use of two, four, or more line wires or communication channels between control and substation.

A coded train of impulses totalling to a constant number is used for the common-diagram system, giving the numerical constancy of the impulse trains a simple and reliable safeguard against errors. For control selection, 91 codes are provided from 15 impulses, divided into three digits. A selection having been made, further final "operate" impulses complete the closing or tripping desired, as well as indicating back immediately that this has been done.

For the switchgear position-indication signalling a fixed number of impulses is again used, but with the difference that each impulse indicates by its polarity either the closed or open position of a particular switch, the fixed total number of the pulses again providing a ready means of checking that no mutilation of impulsing has occurred.

Two pilot lines only are required between the substation and control station equipments for these circuits, the nature of the currents transmitted over them being as follows:

A. *For Control Selection.* A coded train of negative impulses of constant length.

B. *For Final Operation.* A series of dovetailing impulses, transmitted alternately from each station and of positive or negative polarity corresponding to the closed or open condition of the switchgear selected, the actual operating pulse to change this condition being made longer than the others.

C. *For Position Indication.* A fixed-length train of impulses, each impulse of positive or negative polarity corresponding to a closed or open position of an associated switch.

D. *For Line Proving.* A small direct current, removed from the line during impulsing.

3.2 OUTLINE CIRCUIT DESCRIPTION

Fig. 5 is a composite illustration showing association between the circuits.

In the basic supervisory system, a symmetry exists between the circuits at the substation and the control station, because both have to send and receive impulses, the substation to select indicating lamps, etc., at the control station, and the latter to select closing and trip-interposing relays at the substation. A sender-connector circuit at each station enables the selecting impulses to be transmitted from any of a number of control selection keys or switchgear auxiliary contacts without confusion, whilst either indication relays or selector circuit relays enable the received codes to be operative at their respective locations.

Relays connected in the line-switching and impulse circuit enable the two equipments mutually to interlock to avoid confusion whilst impulsing, and also to effect switching between the various circuits.

Given a basic system of supervisory remote control, the application of this to a common-diagram system can now be described in outline form under the two stages into which it naturally divides—establishing connection with a substation, and dressing the common diagram to display that substation.

3.3 Signalling to Establish Pilot-Line Connection from Substation via Tandem to Control Station

Both the control engineer at will, and the substation automatically, must be able to establish this circuit, but whereas in the first case the connection when established may be retained for control purposes, this must not occur when the substation automatically establishes it. If it did, the tandem-control section of the pilots would be prohibited to other substations. The outline circuit is illustrated in Fig. 5.

It is essential that a substation when signalling automatically shall occupy the circuit for the shortest possible time, in order that no delay shall occur to signals from other substations affected by the disturbance. Accordingly, the circuit is so arranged that only the substation number is transmitted by the tandem selector *P* (after it has been actuated from the substation line relay) into the tandem connector *Q*, where a selector is positioned to actuate the substation marker circuit *R* and light the pilot lamp on the system diagram. As soon as the substation marker circuit individual to the calling substation

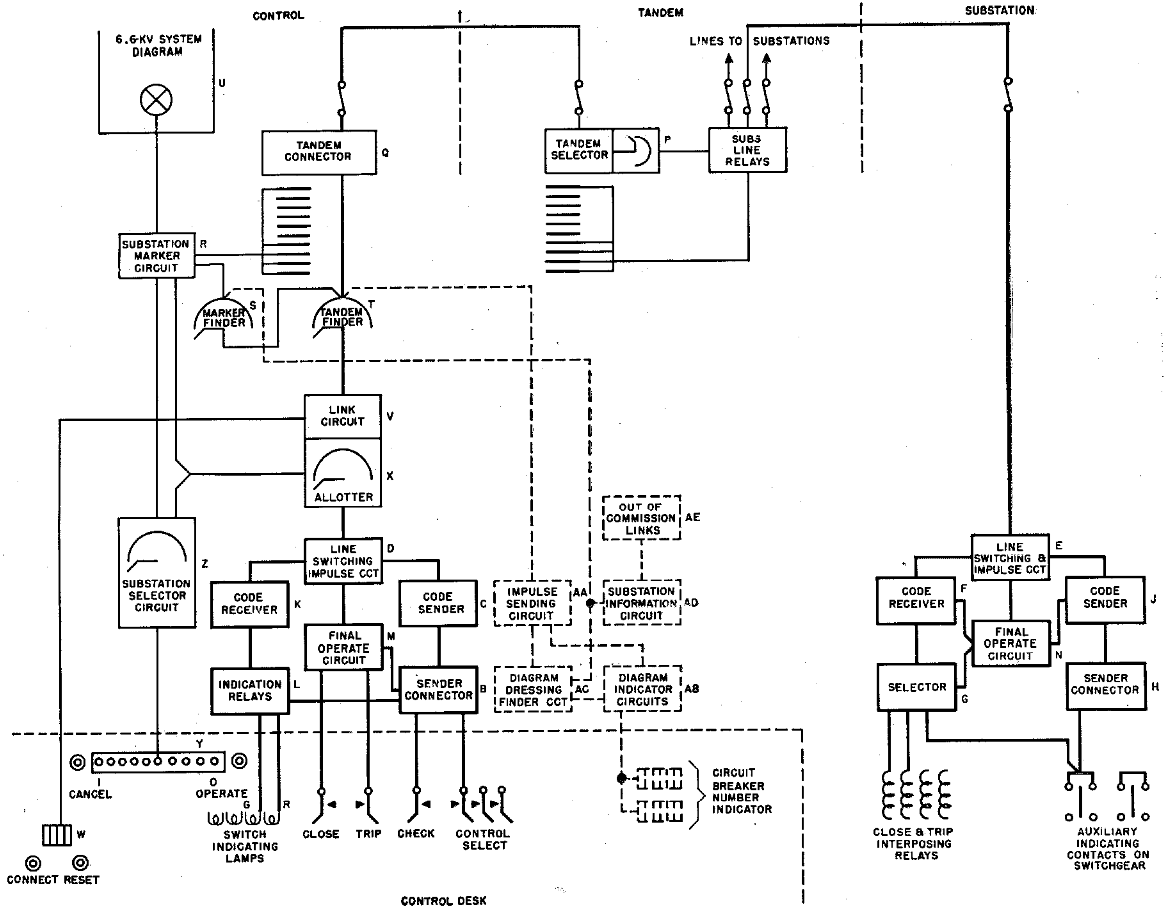


Fig. 5—Composite key diagram. The heavy line indicates the basic supervisory system, the light line illustrates the circuit to establish pilot-line connection from common-control through tandems to substations, and the dotted line the circuit for dressing the common diagram.

has been actuated, the circuits *P* and *Q* are free to operate for other substations, and therefore as many pilot lamps as necessary can be energized in sequence. In addition to the substation pilot lamps, five link circuits *V*, each with a number indicator *W* mounted on the desk, are employed to display the numbers of the first five substations automatically signalling-in a network disturbance. These link circuits are allocated for use to the incoming signals and the control engineer by the allotter circuit *X*, which causes the allocated link circuit to position the number indicator via the markings set up on the tandem finder *T* and marker finder *S*. When the number of a substation is displayed, the system-diagram pilot lamp changes from flashing to steady light, thereby informing the control engineer of this fact.

To "call" a substation, the control engineer must connect a free link circuit to the common diagram and then originate a four-figure number, the first two numerals of which nominate the tandem station to which the required substation (represented by the last two numerals) is connected. Ten digit keys *Y* enable him to do this and store the number in the substation selector circuit *Z*, which becomes effective via the circuits *R*, *S*, and *T*, and light the substation pilot lamp on the system diagram, after he has pressed an "operate" key.

By this arrangement the equipment is set up ready to call the substation, but the pilot-line circuits are not occupied until the control engineer proceeds to check and control the substation.

3.4 Dressing the Common Diagram

Whenever the control engineer wishes to record or control switch positions at any substation, he requires the common diagram dressed to that particular substation. Each possible connection to the busbars is represented on the common diagram by a number indicator, control-select key, and indicating lamps. In a few cases where feeders are solidly connected to the busbars, the control-select key and indicating lamps are omitted.

Setting the number indicators to display, for example, F215 representing Feeder 215, or T4 for Transformer 4, is the principal function of the circuit shown in broken lines in Fig. 5. The range of the information provided by the number indicator is given in Table I, whilst the illustration in Section 5.4 shows an indicator from which the cover has been removed.

Each wheel of the number indicator is moved independently, and therefore to set all 6 wheels of an indicator completely requires 6 trains of up to 10 impulses; these are delivered simultaneously to all wheels at a stepping speed of 10 impulses per second. A common source provides the impulses for all circuits; this arrangement means that, for any particular setting, each wheel has only to be provided with a circuit marking how many impulses are to be supplied to it.

Employing uniselector switches each with 8 levels of 25 points each, one such uniselector circuit or diagram-dressing finder can mark an indicator for 25 different 6-digit positions, the other two levels being available for positioning the uniselector.

In this way a group or rank of, say, 10 diagram-dressing finders, one per indicator, will enable the diagram to be dressed to show 25 different substations of up to 10-switch capacity.

It was found possible to provide for a theoretical number of 150 substations in 6 such ranks for Manchester Corporation, the smallest positioning 4 indicators and the largest 30. As additional substations are connected it may be found necessary to add additional ranks just before the 150-substation limit is reached, unless the sizes of substations are such that they can be fitted into the spaces still available on the existing ranks.

In the key diagram (Fig. 5), the outline of the circuits for dressing the diagram is shown. A

TABLE I

Position	1st wheel	2nd wheel	3rd wheel	4th wheel	5th wheel	6th wheel
1	F	1	1	1	1	a
2	M	2	2	2	2	b
3	T	3	3	3	3	c
4	R	4	4	4	4	d
5	C	5	5	5	5	e
6	—	6	6	6	6	f
7	—	7	7	7	7	g
8	—	8	8	8	8	h
9	—	9	9	9	9	i
0	—	0	0	0	0	j

F = Feeder.
M = Motor convertor.
T = Transformer.
R = Rectifier.
C = Consumer.
a = Incoming.
b = Outgoing.
c = Outgoing.

common source of impulsing is provided in an impulse-sending circuit *AA*, which provides any number of impulses between 1 and 10 to as many indicator circuits *AB* as the diagram-dressing circuit *AC* orders. The source of impulses at 10 per second is a pendulum relay, which operates slave relays and a uniselector in the impulse-sending circuit. The slave relays pass the impulses to each indicator wheel, the number being controlled from one of 10 wires feeding 1 to 10 impulses according to the wiring connections of the diagram-dressing finders, while the uniselector apportions the impulses on to the 10 wires.

One other circuit is engaged in the operation of dressing the diagram. This is the substation information circuit *AD*, which supplies the following information.

The control circuit is informed of the number of signalling groups which will be transmitted by the substation, so that the code receiver at the control station may accept this number only. As the substations vary in size from one switch up to 28, the substation equipments were made so that they could be equipped to transmit information in a maximum of three groups of impulses using 1, 2, or 3 groups as necessary.

If the substation is partially or wholly out of commission, this fact is indicated on an alarm panel on the control desk, and can be manually set up on a separate panel (shown in the key diagram at *AE*) in the control room by inserting plugs in jacks associated with the individual substations.

The interconnector circuits and busbar couplers are the only circuits on the common diagram fitted without indicators, since their purpose is obvious, and consequently substations fitted with interconnectors and/or busbar couplers have to be signalled as such to the common supervisory control equipment by the substation information circuit.

4. *Outline of Operational Procedure*

4.1 SUPERVISORY REMOTE-CONTROL SIGNALLING BETWEEN CONTROL AND THE SELECTED SUBSTATION

A substation having been selected for display and the diagram dressed to illustrate its particular arrangement on the common diagram, the control engineer has only to press the "check" key to cause the display of all existing switch positions and the condition of the alarm circuits. This "check" key is mounted with the other common "operate" keys on the desk, and is connected to the sender connector *B* of the control equipment, which is already connected through the link circuit *V*, and the tandem finder *T*, to the tandem connector *Q*. Its action causes the pilot lines, from the tandem station associated with the substation, to be switched through, and to start the code-sender circuit. This originates a code, the first part of which is the number of the substation, transmitted as three digits totalling 21 impulses, while the second part is a check pulse. The first part of this double code is received at the tandem station and selects the pilot lines to the required substation, and after this first part has been correctly received and has connected the lines through at the tandem station, the second part of the code is transmitted direct into the substation equipment.

A check signal is immediately effective, its action being to cause all the indication signals to be returned from the substation direct into the code receiver. On the completion of these signals the pilot lines are freed for use by any other station.

Indication signals from the substation of all two-position devices are given in 1, 2, or 3 groups, depending upon the size of the substation, the details of each group being shown in Table II.

TABLE II

Group 1		Group 2		Group 3	
Buchholz (surge)		BC 2	Busbar selection and breaker position	OCB 15	Busbar selection and breaker position
Buchholz (gas)		IC 1		OCB 16	
Distribution fuse		IC 2		OCB 17	
Earth-fault indication and trip		OCB 7		OCB 18	
Earth-fault indication		OCB 8		OCB 19	
Fire		OCB 9		OCB 20	
Spare alarm		OCB 10		OCB 21	
OCB 1	Busbar selection and breaker position	OCB 11		OCB 22	
OCB 2		OCB 12		OCB 23	
OCB 3		OCB 13		OCB 24	
OCB 4		OCB 14		Spare OCB	
OCB 5		D 1	Lock-out indications		
OCB 6		D 2			
BC 1		D 3			
2 spare OCB		D 4			
		D 5			

The substation information circuit has already indicated to the control circuit how many groups to expect and, as the signal codes are received from the substation and checked for correctness, the indicating lamps on the control desk are energized in accordance with the signal, i.e., either a red or green lamp for each breaker, one of two white lamps to indicate to which busbar a breaker is connected, or a lamp beneath a glass panel, showing the type of alarm for an alarm condition. These signals are given as short positive or negative impulses for conditions which have not recently changed, but for switch-tripping a longer negative impulse is transmitted. Whilst this longer impulse does not prevent the total being correct, it does enable the control circuit to make a special note of the occurrence and indicate that particular switch with a flashing light. The substation equipment is so designed that this recent-change indication can be stored for a matter of 10 minutes, and also that it will be sent only once. By this arrangement, although a substation may send only its number and not its detailed information immediately a switch trips, when it is checked by the control engineer it indicates especially those switches that have tripped.

4.2 Control Operation

Pressing a selection key mounted between the red and green lamps indicating the breaker he requires to control, the control engineer causes the control circuit to send a switch-selection code of 15 impulses divided into 3 digits, following the station selection code of 21 impulses, through the tandem station to the substation, where it selects the switch-closing and switch-tripping coils, and also connects the final "operate" circuit at the substation to the pilot lines. This is signalled to the control engineer by the light in the particular selection key changing to a rapid flash (or flicker), whilst simultaneously the close or trip keys, common to the whole desk, light up red or green. At the same time the final "operate" circuit at control is connected to the lines, so that the two final "operate" circuits interact with interlacing pulses and retain the pilots to the exclusive use of the engineer at one end and the selected switch at the other.

A long positive impulse will be sent if the control engineer presses the close key, and at the substation this will energize the closing interposing contactor. When the switch closes, a long positive pulse returned from the substation will change the green indication to red on the control desk for that switch. Similarly, pressing the trip key will perform the reverse operation, using long negative pulses, and until the control engineer presses the reset key he can close and trip the selected circuit breaker as often as desired. He cannot, however, continually energize the switch-closing circuit by keeping his control key pressed, the circuit being arranged to operate once and once only for each distinct key operation.

4.3 Operating Times

The periods of time required for operations and indications on the common-diagram system indicate the speed with which a large number of substations can be controlled.

A substation signal of a change of conditions takes 4.5 seconds.

Signalling from different tandem stations can occur simultaneously, but two or more substations connected to the same tandem station must signal in sequence, each taking 4.5 seconds.

The largest number of substations connected to any one of the 12 tandem stations is 24; therefore the time taken to signal a complete shutdown of 288 substations would be 1 minute, 48 seconds.

The time required for the complete display of the largest substation giving the positions of 28 breakers, their busbar selections, and 7 alarms, representing a total of 59 indications, is 23.5 seconds.

To select and operate the switch and indicate its operation requires 12.5 seconds.

5. *Equipment*

5.1 Apparatus

Automatic-telephone switching apparatus has been used throughout, and this, with the exception of the pendulum relay, combined key and lamp, and the number indicator, conforms to standard Post Office designs.

5.2 Pendulum Relay

To provide uniform timing of impulses, pendulum-controlled impulse relays are used as the fundamental source of uniformly timed impulses. The relays are similar to the P.O. 3000-type major relay, but are provided with a horseshoe extension at the end of the relay core and a vibrating reed operating within the arms of the horseshoe. A pair of contacts on each side of the operating reed, mounted in a somewhat similar manner to those of the normal relay, are operated by the reed at its fundamental frequency, one of the contacts being used to drive the reed through the medium of the energizing winding on the core.

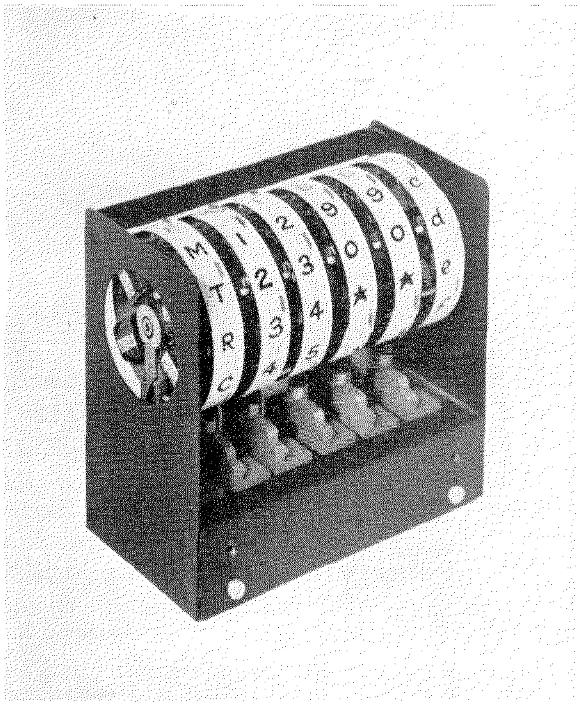
5.3 Combined Key and Lamp

To reduce the amount of control-panel area, a control key which includes an inset indicating lamp has been developed. This comprises a diecast body, on the sides of which are mounted twin-contact lever springs of the type used for automatic-telephone relays. Inside the body and between the springs is carried the lamp mounting, which projects into the barrel of the operating knob carried in the front of the die castings. This barrel is shaped and spring-loaded so as to provide a number of different types of movement of the knob. It can be made to operate the springs by turning through an angle of 90 degrees; by a push-and-pull action; by a push-in spring-return action; or by a short spring-return action on either of two positions normally 90 degrees apart.

It is arranged for single-hole fixing in the control panel, and the operating knob is 1 inch in diameter with an indicating window approximately 1 inch in diameter. The indicating window is therefore quite large enough for the display of an abbreviated description of the function of the key, whilst arrows inscribed in the window can be used to indicate whether the key has to be turned or pressed.

5.4 Number Indicator

To designate substations, feeders, and transformers, electrically operated indicators have been developed, built on the unit principle, in which each wheel, showing one digit of a number, with its operating mechanisms, forms a complete and separate unit. In this respect it is quite unlike the better-known Veeder counter and has the great advantage that all wheels can be operated simultaneously and independently to any of their positions (see illustration).



Number indicator with cover removed.

Each unit comprises a simple 4-pole stator and a wheel-type rotor with bent-over fins revolving about the centre of the stator and around the outside of the stator poles. By energizing the stator poles in pairs, the rotor is attracted to move through 30 degrees, in two movements of 15 degrees each, against a light return spring. A small spring catch beside each pole of the stator is energized by the operating windings and prevents backward movement of the rotor; it thereby ensures continuous forward movement so long as impulses are delivered to the pair of stator windings. The indicator windings remain energized for as long as the display is required, and for this purpose an external resistance is introduced by the control circuit to economize in current and reduce heating. The digits appearing in the instrument window are quarter-inch characters for capitals, in black on a white background.

By using the number indicator, a designation comprising 6 characters is displayed in a length of 3 inches, the whole indicator in its casing being approximately 4 inches long by 2½ inches wide by 3½ inches deep. The indicators are wired to a removable multi-point plug to facilitate maintenance.

5.5 SUBSTATION EQUIPMENT

At each substation, the cubicle containing the relays and switches is one of five standard sizes. Equipment external to the cubicle consists of a battery, with a metal-rectifier automatic charging supply, and interposing relays which are mounted upon the panels or cubicles of the switchgear to be controlled.

With the exception of the largest, the supervisory cubicles were designed for wall-mounting with front access only, the relays and switches being mounted upon a drop frame, hinged along its lower edge, to provide the necessary access to the wiring. The relays, suitably grouped into circuits, were mounted upon removable jack-in panels. In the largest substation cubicle, front and rear access was provided to enable the terminal units for the interconnecting cables between the supervisory equipment and the switchgear to be accommodated on each side of the central framework.

The question of unit-type design to facilitate standardized arrangements for the wide variations of requirements is of considerable importance, if maintenance and spares are to be reduced to a minimum. The requirements vary between one switch indicated only, plus alarms, to 28 switches controlled and indicated, the number of units of equipment used being given in Table III.

TABLE III

Facilities and equipment	Capacity of cubicle				
	A	B	C	D	E
Indicated OCB's	4	10	8	12	28
Controlled OCB's	—	—	8	12	28
Alarms	7	7	7	7	7
Telephone	—	4	—	—	—
D-C lock-outs	—	—	5	5	5
Number of indicating groups	1	2	2	2	3
Relay panel (design 1)	1	1	1	1	1
Relay panel (design 2)	1	1	1	1	1
Relay panel (design 3)	—	—	1	1	1
Uniselectors	1	1	3	4	5

5.6 CONTROL-STATION RELAY-ROOM EQUIPMENT

In the relay room for both the common-diagram system, and eight 33-kilovolt substations, the operation of which has now been centralized in the control room, provision has been made for 8 rows of racks, only 5 of which are occupied at present. The whole of the relay and switching equipment for the common-diagram system occupies 3 racks at one end of the room, and cabling has been initially installed for 400 substations. The present floor space will not be exceeded by the ultimate expansion planned.

At the other end of the room two suites are occupied by the supervisory relay racks for the 33-kilovolt substations which were remotely controlled from other centres, previous to the planning of the present scheme. The central space at present unoccupied, approximately 15 feet by 21 feet, is considered ample for any future developments. Racks 8 feet, 6 inches high and either 2 feet, 6 inches or 4 feet, 6 inches wide, depending upon the equipment to be mounted upon them, have been used for the common-diagram system. Equipment is mounted upon one side of the rack, with wiring on the other side, and, wherever convenient and necessary for testing or maintenance, relays and switches are mounted upon jack-in panels. A cross-connecting frame enables all wiring between the substation marker and finder circuits and the diagram-dressing circuit, which is particular to the substation, to be readily provided as further substations are brought into use.

5.7 CONTROL ROOM

Particulars of the supply-system control room are shown in Fig. 6. The batteries, charging gear, pilot-cable terminals, supervisory-switch, and relay equipment are on the ground floor.

The upper floor houses ventilating and heating plant, a telephone exchange, canteen, cloakroom, log store, system diagrams, and control desk.

The control room is elliptical, the walls being formed of panels for supervisory control of the 33-kilovolt equipment, lighting and ventilating-plant switchgear, meters, load-summation indicators, and complete system diagrams. A ceiling specially designed by illumination experts was provided to accommodate coloured fluorescent lighting tubes, which give a daylight effect, a high standard of illumination on horizontal and vertical surfaces, and freedom from shadow and glare.

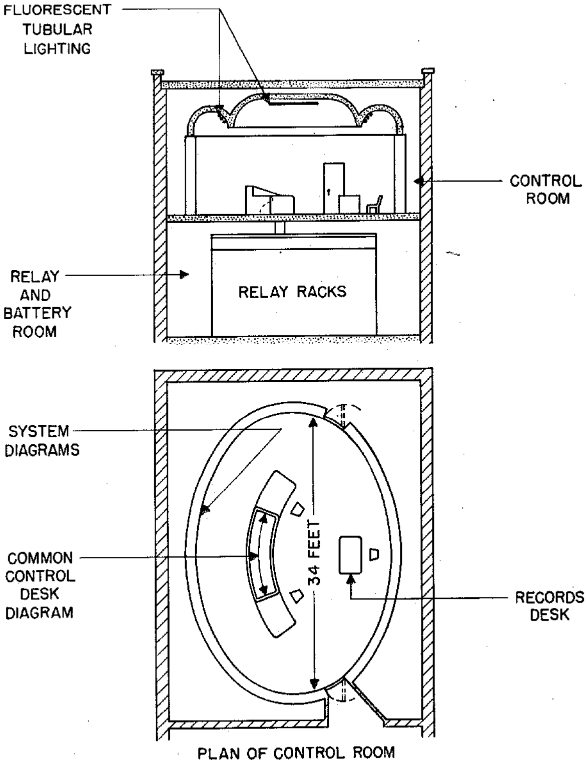


Fig. 6—General arrangement of a control room housing a common-diagram system.

In the middle of the room, the control desk has accommodation on the wings for two control engineers, Post Office telephone and private telephone facilities for communication with every staff office, generating station, substation, and depot on the system. There is also a card index of substation data, diagrams, names, and code numbers. The common-diagram control equipment occupies the middle section of the desk.

The panel walls are coloured *eau-de-nil*, the beading, doors, and furniture mahogany, and the floor pink with a dark green and black border.

6. Considerations of Cost and Space

6.1 COSTS

Considerable economy would be expected both in first cost and when extending a common-diagram system, as compared with individual supervisory equipment for each substation. The cost of supervisory control equipment for a substation may be divided into two parts, (*A*) control-station equipment, including the control diagram board, and (*B*) substation equipment, including interposing relays and cabling. Both costs may be sub-divided into a basic part plus an adjustment for each switch-control indicator and alarm. In general, the basic equipment is designed to cater for a maximum number of facilities and will vary accordingly. It is rarely fully equipped at the outset, being intended to cater for future extensions within the forecast maximum.

In addition to these normal variations, there are other considerations to be taken into account when assessing the probable cost of installation. These concern the degree to which the control station is to be completely equipped so as to be agreeable to and maintain the alertness of its occupants; the cost of pilot cables necessary; modifications of switchgear to add electrical closing and tripping gear; and whether or not battery equipment additional to that already existing is to be provided.

Some form of cost comparison, however, can be obtained by assuming either that supervisory equipment only is to be supplied, or that supervisory equipment plus interposing relays, cable, battery equipment, and installation, are included. The following comparisons have been made upon the basis of fully equipped stations, each with 20 switches controlled and indicated, plus five alarms indicated only.

It might be supposed that with the common-diagram system there is a minimum number of substations at which it becomes economic when compared with individual equipment for each substation. This is correct, but the minimum number varies with the ultimate capacity of the common-diagram system, since it is obvious that the economy effected is due almost wholly to a saving in control-station equipment, and also that the amount of common equipment provided for a common-diagram system with an ultimate capacity of 25 substations will be smaller than that provided for 50, 100, 200, or more substations.

It should be noted that for an individual equipment providing only for indication and alarms, the substation section cost will be between 35 and 45 per cent of that of the whole equipment, whilst if equipment for both control and indication is provided, the substation cost rises to between 55 and 60 per cent of the whole cost. This increase is indicative of the high proportion of the whole cost required to provide for interposing relays and cabling.

In Fig. 7, curve *A* shows the minimum number of substations, expressed as a percentage of the maximum substation capacity of the system, at which the common-diagram system will prove economic when the substations are fully equipped. Curve *C* indicates the percentage saving at first cost effected when considering a fully equipped common-diagram system, as compared with individual systems to give the same facilities. For comparison, curve *B* indicates the percentage saving of first cost of the supervisory equipment only of a common-diagram system compared with individual systems, and shows a higher saving than that indicated in curve *C* for the same equipment complete with all cabling, accessories, and power supplies installed ready for service. Both curves exclude any cost for either switchgear modifications or pilot lines.

6.2 Space Economy

Equally as important as economy of cost is the saving in control building-space achieved by the adoption of a common-diagram system and it is possible to assess the approximate space saved on a hypothetical basis. For this purpose, we need only consider the control room and its associated relay room, since the substation requirements will be approximately the same for individual systems as for a common-diagram system. The following curves are based upon substations equipped to control and indicate 20 circuit breakers and five alarms.

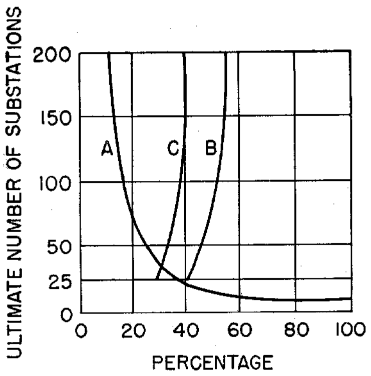


Fig. 7— *A*—Minimum number of substations equipped initially for which common-diagram system is economical compared with individual systems, expressed as a percentage of the number to be equipped ultimately. *B*—Saving in first cost of common-diagram-system supervisory equipment only, expressed as a percentage of equivalent individual systems. *C*—As *B*, but including costs of cabling, interposing relays, power supplies, erection, and testing.

Curves *A* and *B* in Fig. 8 illustrate the difference in relay room space for the two systems, and Fig. 9 compares the control-room space for individual and common-diagram system requirements. These curves can, of course, only be approximate and are based on the following estimate:

A. That an individual system of the size stated would require 5 square feet of control diagram per substation.

B. That the individual control panels, each with a maximum height of 6 feet operating space, would be arranged in a semicircle for up to 25 substations (ultimate), and in a circle for a larger number of substations; also that a minimum control-room size of 25 feet diameter for the semicircular pattern would be chosen. The curves are then based on areas semicircular or circular, adding nothing for the corners which usually result when a circular room is placed within a rectangular building.

C. For the relay room, rack space for the individual system is assumed to be 4 feet high by 3 feet wide per substation, using racks 8 feet, 6 inches high. No headroom allowance has been made for the end of suites.

D. Space in common-diagram system calculations is of the same order as that provided at Manchester.

Fig. 10 indicates the total percentage saving in space obtained at the control centre on these hypotheses. As would be expected, the savings increase with the number of substations controlled, attaining figures showing an economy of 86 per cent in space for the largest installations.

6.3 Maintenance Costs

Records of cleaning and maintenance costs for individual-type supervisory equipment show that the amount per substation is small, being almost entirely for labour.

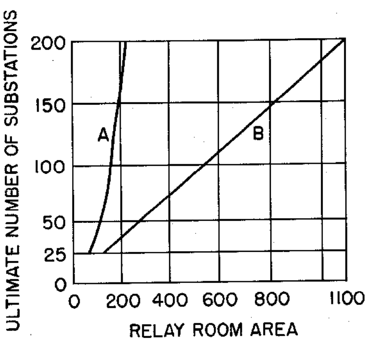


Fig. 8—Comparison of relay-room area, *A*—common-diagram system, and *B*—equivalent individual system.

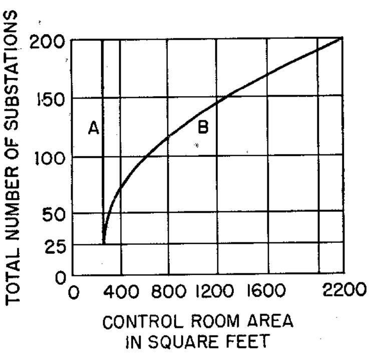


Fig. 9—Comparison of control-room area, *A*—common-diagram system, and *B*—equivalent individual system.

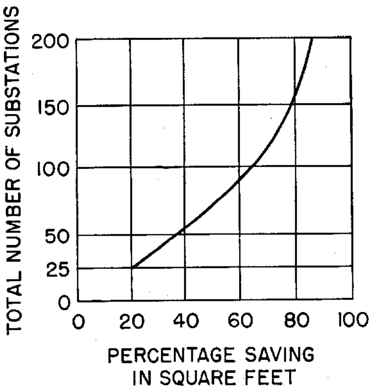


Fig. 10—Total saving in space obtained from comparisons shown in Figs. 8 and 9, expressed as a percentage of space required for individual systems.

The relatively smaller amount of equipment required at the control station where a common-diagram system is installed, and also the large degree of standardization obtained for the relays, switches, etc., would indicate that the maintenance costs would be *pro rata* on a quantity basis, since similar equipment is used; consequently these costs should show a percentage economy as compared with the individual system at least equal to that shown in Fig. 7, curve *C*, for the economy of first cost.

7. *Conclusion and Acknowledgments*

It is thought that, by the development of these ideas, the use of supervisory remote-control equipment has been extended on normal automatic-telephone switching practice to a wider application than was practicable with individual systems for large numbers of substations. As might be expected, this results in an economy of cost of equipment, an economy of cost for control buildings, and an economy of effort of the control staff. Operational control of large electricity supply systems will undoubtedly produce new requirements and problems as industrial, traction, and domestic loads increase, but it is believed that, in the system described, a basis for meeting these requirements without adversely affecting present facilities has been provided.

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Anomalous Attenuation in Waveguides*

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THE puzzling phenomenon of decreasing attenuation constant with increasing frequency which occurs in a few isolated instances is here elucidated by treating the guides concerned as limiting cases of a guide of more general shape in the interior of which the waves display the normal properties characteristic of waves in guides generally. The equations of the electromagnetic field, cut-off frequency, and attenuation constant describing the isolated cases are then, in like manner, deduced as limiting cases from those appropriate to a guide of general shape. The isolated cases thus lose their character of isolation and assume that of straightforward limits instead. According to the point of view developed in the paper, these limiting cases imply an electromagnetic field which extends to infinity along one of the transverse co-ordinates but, being wrapped around the axis of the guide, the field is constrained to exist in finite space where it continues to display the properties characteristic of a field of infinite extent.

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It is a well-known fact that when an electromagnetic wave is traversing a transmission line of any conventional type, it is always possible to find a frequency above which the attenuation constant of the wave is a steadily increasing function of frequency. This property also appertains to waves propagated through the interior of hollow metal tubes with the exception, however, of a few seemingly isolated cases. There is, for instance, the case where the *H*-wave (also known as the transverse electric or *TE*-wave) of zero order is propagated through a tube of circular cross-section. Here, as in two further cases to be dealt with below, the attenuation constant follows a course precisely opposite to that normally encountered; that is to say, in each case a frequency can be found above which the attenuation steadily *decreases* and ultimately vanishes altogether.

This striking anomaly may be accounted for satisfactorily by means of Ampère's Law of electromagnetic induction, whereby the current induced in the metal tube may be stated in terms of the tangential components of the magnetic field at the surface of the metal—the longitudinal component of the current in terms of the transverse component of the field—and the transverse component of the current in terms of the longitudinal component of the field. Now in the exceptional case of the *H*-wave of zero order in a guide of circular cross-section, all lines of magnetic force are at any frequency restricted to radial planes (Fig. 1) and in consequence the transverse tangential component is always zero. There thus remains only the longitudinal tangential component and as this component—in common with the longitudinal component of all other waves in hollow metal tubes—is a steadily decreasing function of frequency, the attenuation of the wave is of necessity also a diminishing function of frequency.

As an instructive alternative explanation it is here suggested that a guide of circular cross-section might be regarded as an extreme case of a guide having a cross-section of sector shape (Fig. 2). In such a guide the *H*-wave of zero order displays the normal properties characteristic of waves in guides generally and from these there emerge as straightforward limiting cases those appropriate to waves in a circular guide. By this procedure, the appearance of an anomaly is completely eliminated. Moreover, by extending the analysis to guides encompassed by two coaxial circular cylinders and two radial planes (Fig. 3), the properties of *H*-waves of zero order may be stated in a general form which includes as particular cases those associated with guides of rectangular, circular, sector- and ring-shaped cross-sections, with and without baffle planes.

We shall first recapitulate for ready reference the well-known relevant formulae relative to a

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rectangular guide. We shall then establish the corresponding formulae for a guide of sector shape leading to those for circular guides. Finally we shall proceed to the general case and identify as particular cases the results recorded for the guides of more special shape.

1. Principal Symbols

a = width of rectangle in metres

b = height of rectangle in metres

r_0 = radius of sector in metres

r_1, r_2 = inner and outer radii of ring-sector in metres

ϕ = sector angle in radians

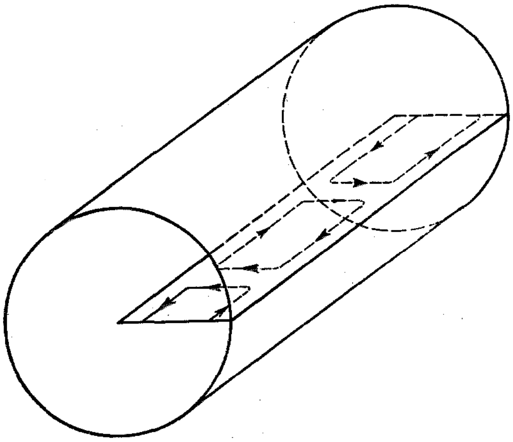

Fig. 1—Pattern representing the magnetic field in any radial plane of a circular guide transmitting an *H*-wave, of zero order and principal mode. If the plane is turned full circle the pattern describes a set of toroidal surfaces which completely represent the magnetic field in space.

$\mu, \bar{\mu}$ = permeability of dielectric and conductor, respectively, in henrys per metre; for vacuum $\mu_0 = 4\pi \times 10^{-7}$

$\epsilon, \bar{\epsilon}$ = permittivity of dielectric and conductor, respectively, in farads per metre; for vacuum, $\epsilon_0 = \frac{1}{36\pi} \times 10^{-9}$

$g, \bar{g}$ = conductivity of dielectric and conductor, respectively, in mhos per metre; for pure copper $\bar{g} = 5.8 \times 10^7$

$c = \frac{1}{\sqrt{\mu\epsilon}}$ = characteristic velocity of waves in the unrestricted medium in metres per second; for vacuum $c = 3 \times 10^8$

f = frequency in cycles per second

f_{0m} = cut-off frequency of the mth mode in cycles per second

$\nu_{0m} = \frac{f_{0m}}{f}$

$\omega = 2\pi f$

α_{0m} = attenuation constant corresponding to the mth mode in nepers per metre

Γ = propagation constant in the z direction per metre

k_{1m} = mth non-vanishing root of $J_1(x) = 0$; i.e., $k_{11} = 3.83$; $k_{12} = 7.02$; $k_{13} = 10.17$.

2. Guides of Rectangular Cross-Section

The electric and magnetic intensities of an H-wave of zero order and mth mode[1] in a guide of rectangular cross-section, having metal walls of infinite

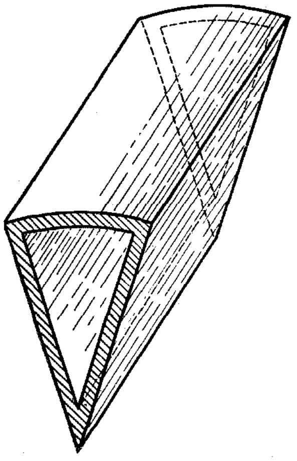

Fig. 2—Guide of sector-shaped cross-section.

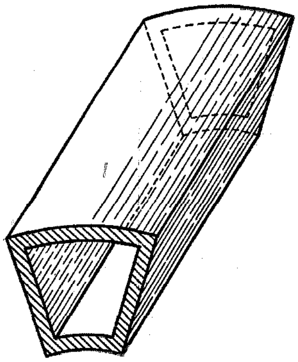

Fig. 3—Guide composed of portions of coaxial circular cylinders and radial planes.

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[1] The characteristic feature of any *H*-wave is a prominent magnetic field in the direction of propagation. To rank as a wave of "zero" order it is required that in the direction of one of the transverse co-ordinates of the guide—for instance, the height of a rectangular guide—every one of the six components of the electromagnetic field is independent of the co-ordinate, that is to say, is of "zero variation" in that direction. The mode m specifies the number of half-cycles to which the non-vanishing components of the field are subjected in the direction of the other transverse co-ordinate of the guide—for instance, the width of a rectangular guide. The first mode is frequently referred to as the principal mode.

conductivity and a dielectric that is nondissipative, are given by

$$\left.\begin{aligned} &E_x=0, \qquad E_z=0, \qquad H_y=0,\\ &E_y=-\frac{m\pi}{a}\sin\left(\frac{m\pi}{a}x\right)e^{-\Gamma z},\\ &H_x=\left(\frac{1-\nu_{0m}^2}{\mu/\epsilon}\right)^{\frac{1}{2}}\frac{m\pi}{a}\sin\left(\frac{m\pi}{a}x\right)e^{-\Gamma z},\\ &H_z=\frac{1}{j\omega\mu}\left(\frac{m\pi}{a}\right)^2\cos\left(\frac{m\pi}{a}x\right)e^{-\Gamma z}. \end{aligned}\right\} \quad (1)$$

If the conductivity of the metal walls is finite but high, the above expressions of the field become approximations but remain amply accurate for our purpose. As usual the time factor exp $j\omega t$ is omitted.

Since both E_x and E_z are zero, the lines of electric force may be represented by patterns of parallel lines stretching from top to bottom with densities that vary sinusoidally across any transverse section of the guide. For the case where $m=1$ (Fig. 4), the electric field is a maximum along the middle line of the section and zero at each of the two side walls. This pattern is subject to the usual cyclic variation with time, that is to say, the pattern fades away, increases in the opposite direction to its original intensity, fades again, and finally re-assumes the original configuration from which the next cycle begins.

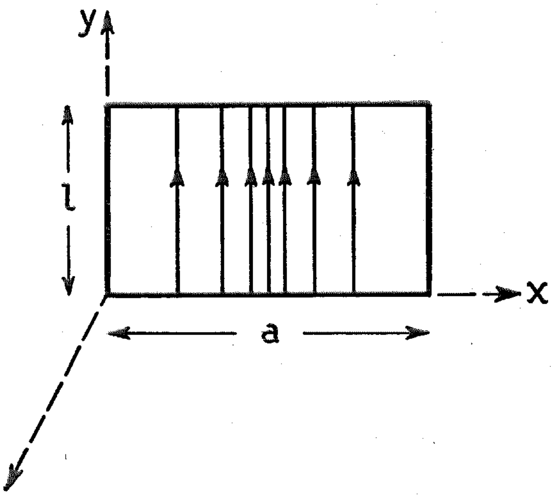

Fig. 4—Pattern representing the electric field of an *H*-wave, of zero order and principal mode, propagated through a rectangular guide.

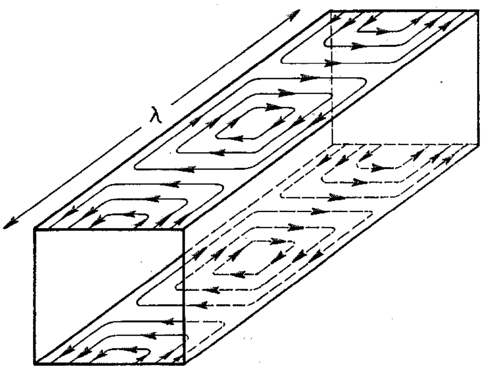

Fig. 5—Pattern representing the magnetic field in any horizontal plane of a rectangular guide transmitting an *H*-wave, of zero order and principal mode. λ indicates the length of the wave.

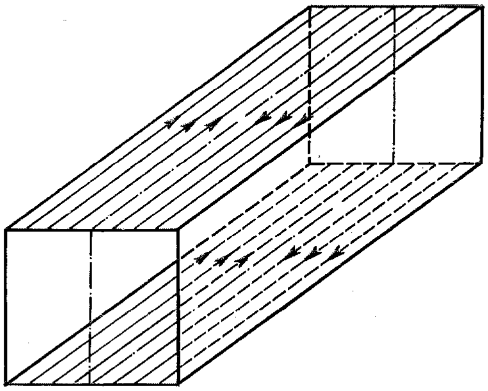

Fig. 6—Pattern representing the magnetic field in any horizontal plane of a rectangular guide supporting an *H*-wave, of zero order and principal mode, at the cut-off frequency.

The lines of magnetic force are restricted to horizontal planes and, since both H_x and H_z are independent of the height of the guide, the magnetic pattern is identical in all horizontal planes. For $m=1$ (Fig. 5), there is a single row of closed curves. The distance between recurring configurations of the pattern indicates the wavelength. With falling frequency, the loops lengthen and finally, at cut-off frequency, they break up into two sets of parallel lines as shown in Fig. 6. The magnetic field is now divided into two halves pointing in *opposite* directions and being separated from one another by the vertical mid-plane (shown dotted in Fig. 6) where the magnetic intensity is permanently zero. While the magnetic field is thus being broken up, no corresponding change takes place in the electric field. This field retains at cut-off frequency the character indicated in Fig. 4; that is, at any given instant the electric intensities in both halves of the guide point in the *same* direction. In consequence we are led to the conclusion that energy is simultaneously thrust from the interior towards each of the side walls whence it is reflected back towards the mid-plane. There are thus two transverse pulsations of energy, between mid-plane and side walls, without any concurrent transmission of energy along the guide. The frequency at which this resonant state is reached is given by

$$f_{0m}=\frac{m\pi}{2\pi a}c. \qquad (2)$$

Assuming that the dielectric medium is non-dissipative, as in the case of air, we have for the attenuation constant of the wave

$$\alpha_{0m}=\frac{\left(\frac{\bar{\mu}\pi f}{g}\right)^{\frac{1}{2}}}{\left(\frac{\mu}{\epsilon}\right)^{\frac{1}{2}}}\left[\frac{1}{b}\frac{1}{(1-\nu_{0m}^2)^{\frac{1}{2}}}+\frac{2}{a}\frac{\nu_{0m}^2}{(1-\nu_{0m}^2)^{\frac{1}{2}}}\right]. \qquad (3)$$

The first term is due to currents induced in the top and bottom walls by the tangential magnetic intensities H_x and H_z. For all frequencies above $3^{\frac{1}{2}}f_{0m}$, this term is an increasing function of frequency (Fig. 7). The second term arises from currents induced in the side walls. These are due to the longitudinal magnetic field H_z only; for H_x, being perpendicular to the side walls, is prevented from setting up currents in them. This second term is a decreasing function of frequency (Fig. 7). As the cut-off frequency (2) is independent of b, the height of the guide may be increased indefinitely without affecting the character of the wave. At the same time the first term of the attenuation constant (3) diminishes and ultimately, when the rectangular guide becomes a guide composed of two parallel planes, only the second term of (3) remains; and now, with increasing frequency the attenuation actually tends towards zero. Although this case is of no practical importance it is, as we shall presently see, the exact counterpart of a physically realizable guide, namely the guide of circular cross-section.

3. *Guides of Sector Cross-Section*

In so far as H-waves of zero order are concerned there is a close correlation between a guide of rectangular cross-section in a cartesian system of co-ordinates (x, y, z) and a guide of sector-shaped cross-section in a cylindrical system of co-ordinates (r, ϕ, z) (Fig. 8). The form of the equations of the electromagnetic field is

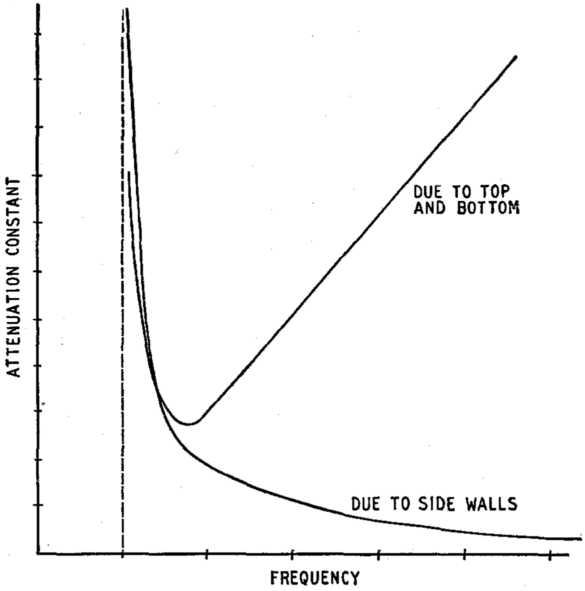

Fig. 7—Analysis of the attenuation constant of an H-wave of zero order propagated through a guide of rectangular cross section.

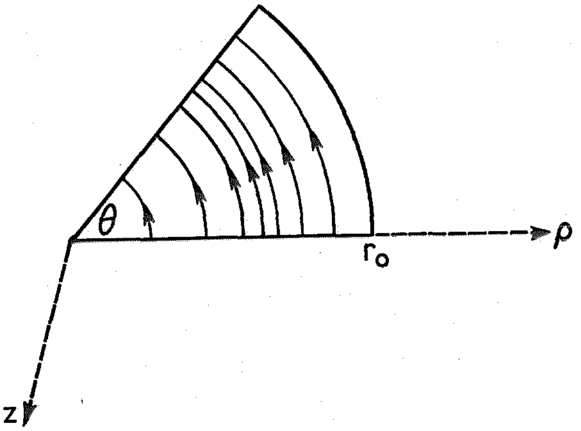

Fig. 8—Pattern representing the electric field of an H-wave, of zero order and principal mode, propagated through a guide composed of a portion of a circular cylinder and two radial planes.

exactly the same in the two cases; in place of the circular functions, we now have Bessel functions of the first kind, and in place of the non-vanishing roots of $\sin x = 0$, we now have the non-vanishing roots of $J_1(x) = 0$, which are denoted below by k_{1m}. Thus

$$\left.\begin{aligned} &E_r = 0, \qquad E_z = 0, \qquad H_\phi = 0,\\ &E_\phi = -\frac{k_{1m}}{r_0} J_1\left(\frac{k_{1m}}{r_0} r\right) e^{-\Gamma z},\\ &H_r = \frac{(1-\nu_{0m}^2)^{\frac{1}{2}}}{(\mu/\epsilon)^{\frac{1}{2}}} \frac{k_{1m}}{r_0} J_1\left(\frac{k_{1m}}{r_0} r\right) e^{-\Gamma z},\\ &H_z = \frac{1}{j\omega\mu}\left(\frac{k_{1m}}{r_0}\right)^2 J_0\left(\frac{k_{1m}}{r_0} r\right) e^{-\Gamma z}. \end{aligned}\right\} \quad (4)$$

The pattern of the electric field now consists of concentric arcs (Fig. 8) and that of the magnetic field of closed curves in radial planes (Fig. 9).

The cut-off frequency is given by

$$f_{0m} = \frac{k_{1m}}{2\pi r_0} c, \quad (5)$$

which differs from the expression found for the rectangular guide merely in that the root of the Bessel function takes the place of that of the corresponding circular function.

As the attenuation constant may be calculated by Schelkunoff's well-known general formula[2] in a perfectly straightforward manner, it may suffice here merely to state the result.

$$\left.\begin{aligned} \alpha_{0m} = \frac{\left(\frac{\mu\pi f}{g}\right)^{\frac{1}{2}}}{\left(\frac{\mu}{\epsilon}\right)^{\frac{1}{2}}} &\left[\frac{2}{\phi r_0} C_{1m} \frac{1}{(1-\nu_{0m}^2)^{\frac{1}{2}}}\right.\\ &\left. + \frac{2}{\phi r_0} \frac{\nu_{0m}^2}{(1-\nu_{0m}^2)^{\frac{1}{2}}} + \frac{1}{r_0} \frac{\nu_{0m}^2}{(1-\nu_{0m}^2)^{\frac{1}{2}}}\right], \end{aligned}\right\} \quad (6)$$

where

$$C_{1m} = \frac{\int_0^{k_{1m}} [J_1(x)]^2 \, dx}{k_{1m}[J_0(k_{1m})]^2}.$$

The factor C_{1m} is always greater than unity. The lowest value occurs for $m = 1$ in which case C_{1m} is approximately 1.01. The first term in (6) arises from currents induced in the radial planes, partly by H_r and partly by H_z. After a minimum is passed, this term is an increasing function of frequency. The second and third terms arise, respectively, from currents in the radial planes and in the cylindrical surface in consequence of the longitudinal field; both of these terms are decreasing functions of frequency. Since the cut-off frequency (5) is independent of the sector angle, (6) is applicable to the case of $\phi = 2\pi$, that is to say, to a circular guide with a single radial baffle plate (Fig. 10).

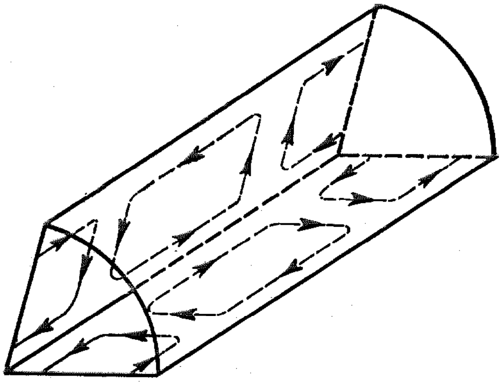

Fig. 9—Patterns representing the magnetic field in any radial plane of a guide of sector-shaped cross-section transmitting an H-wave, of zero order and principal mode.

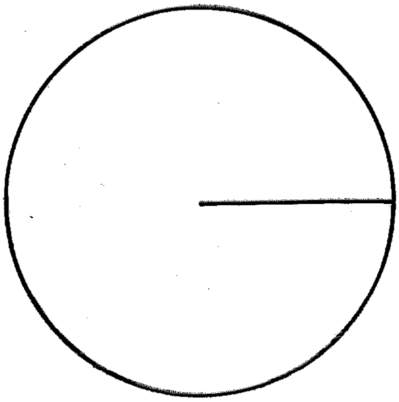

Fig. 10—Guide composed of a circular cylinder and a radial baffle plane.

A sector angle of 2π appears to be the limiting value for a guide in physical space, but this restriction need not be applied to the equations of the field or the equations of the cut-off frequency and attenuation constant. Although originally established for physical space, their range of validity extends into the region of ϕ beyond 2π —a fictitious region, known also as Riemann Space,[3] where the sector angle may, indeed, be increased indefinitely. In consequence, the length of the arc ϕr_0 increases correspondingly, with the result that the first two terms of (6) tend towards zero. As the radial planes of the guide are thus ultimately of no effect they may be removed, and now the barrier between physical and fictitious space effectively breaks down, and there remains a guide of circular cross-section. On this view, then, a guide of circular cross-section is the extreme case of a guide of sector-shaped cross-section, for which the length of the arc becomes infinite—in exact correspondence to the extreme case of the rectangular guide for which the height increases indefinitely. As the first two terms in (6) vanish, the attenuation constant is supplied by the third term alone, which is a falling function with increasing frequency.

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[2] S. A. Schelkunoff, "Transmission Theory of Plane Electromagnetic Waves," *Proceedings of the I. R. E.*, v. 25, p. 1482; November, 1937.

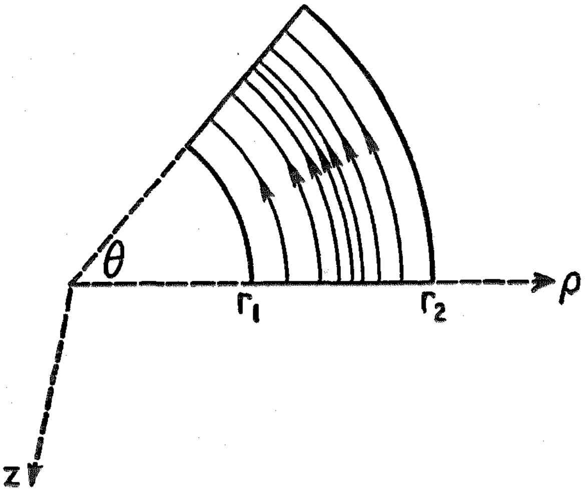

Fig. 11—Pattern representing the electric field of an *H*-wave, of zero order and principal mode, propagated through a guide composed of portions of two coaxial circular cylinders and two radial planes.

As a check of our last result, we may employ a method that does not require the notion of a fictitious space. Let us assume that while the conductivity of the metal of the cylinder retains a high, but finite, value, that of the baffle plane is allowed to increase indefinitely. If then, instead of being given as a product of two factors, (6) is re-written as a sum of three separate terms, the first two terms, which refer to the baffle plane, again tend towards zero. Moreover, as the baffle plane does not constrain the field, the plane may be deleted altogether. There thus again remains a guide of circular cross-section for which the attenuation constant is as before correctly given by the third term of (6).

4. *Guides of Ring-Sector Section*

The treatment of this case (Fig. 11) is identical with that of the sector-shaped guide except that to the Bessel function of the first kind there must now be added that of the second kind. If we put

$$Z_n(\chi r) = AJ_n(\chi r) + BN_n(\chi r),$$

where A and B are real and χ is a function of the radii r_1 and r_2, the electromagnetic field is given by

$$\left.\begin{aligned}
&E_r = 0,\\
&E_\phi = -\frac{\nu_{1m}}{r_2-r_1} Z_1\left(\frac{\nu_{1m}}{r_2-r_1}r\right)e^{-\Gamma z},\\
&E_z = 0,\\
&H_r = \left(\frac{1-\nu_{0m}^2}{\mu/\epsilon}\right)^{\frac{1}{2}} \frac{\nu_{1m}}{r_2-r_1} Z_1\left(\frac{\nu_{1m}}{r_2-r_1}r\right)e^{-\Gamma z},\\
&H_\phi = 0,\\
&H_x = \frac{1}{j\omega\mu}\left(\frac{\nu_{1m}}{r_2-r_1}\right)^2 Z_0\left(\frac{\nu_{1m}}{r_2-r_1}r\right)e^{-\Gamma z},
\end{aligned}\right\} \qquad (7)$$

where ν_{1m} denotes the mth root of the equation

$$J_1(\chi r_1)N_1(\chi r_2) - J_1(\chi r_2)N_1(\chi r_1) = 0;$$

moreover $m\pi \leqslant \nu_{1m} \leqslant k_{1m}$.

The cut-off frequency is given by

$$f_{0m} = \frac{\nu_{1m}}{2\pi(r_2-r_1)}c. \qquad (8)$$

If r_1 approaches zero, (7) and (8) become identical with those appropriate to a sector-shaped guide, that is, (4) and (5). If both r_1 and r_2 are very large, the Bessel functions in (7) may be replaced by their asymptotic expansions whereby (7) assumes the form of (1); i.e., that of the field within a rectangular guide.

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[3] J. H. Jeans, "The Mathematical Theory of Electricity and Magnetism," 5th Edition, p. 283.

As before, the attenuation constant may be calculated by the general formula in a straightforward manner with the result

$$\alpha_{0m}=\frac{\left(\frac{\bar{\mu}\pi f}{\bar{g}}\right)^{\frac{1}{2}}}{\left(\frac{\mu}{\epsilon}\right)^{\frac{1}{2}}}\left\{\frac{2}{\phi}\frac{\int_{r_1}^{r_2}[Z_1(\chi r)]^2dr}{r_2^2[Z_0(\chi r_2)]^2-r_1{}^2[Z_0(\chi r_1)]^2}\times\frac{1}{(1-\nu_{\ m})^{\frac{1}{2}}}+\frac{2}{\phi}\frac{r_2[Z_0(\chi r_2)]^2-r_1[Z_0(\chi r_1)]^2}{r_2^2[Z_0(\chi r_2)]^2-r_1^2[Z_0(\chi r_1)]^2}\frac{\nu_{0m}^2}{(1-\nu_{0m}^2)^{\frac{1}{2}}}+\frac{r_2[Z_0(\chi r_2)]^2+r_1[Z_0(\chi r_1)]^2}{r_2[Z_0(\chi r_2)]^2-r_1^2[Z_0(\chi r_1)]^2}\frac{\nu_{0m}^2}{(1-\nu_{0m}^2)^{\frac{1}{2}}}\right\}. \quad (9)$$

The first term in (9) is due to the currents induced in the two radial planes, partly by H_r and partly by H_z. This is the rising term. The second and third terms are due to currents induced by H_z in the radial planes and in the two cylindrical surfaces, respectively. If r_1 tends to zero, or if both r_1 and r_2 become large, (9) assumes the forms appropriate to guides of sector or rectangular cross-sections. For $\phi=2\pi$, the guide becomes a pair of coaxial cylinders with a single radial baffle plane (Fig. 12). As before, we make use of the notion of Riemann Space and extend the validity of our equations into regions for which the sector angle exceeds 2π. And then with ϕ increasing indefinitely, the first two terms of (9) again tend towards zero; and again the radial planes become ineffective and may be removed. On this view, then, a guide of two coaxial circular cylinders is an extreme case of a guide composed of portions of two coaxial cylinders and two radial planes, and the attenuation constant is simply given by the third term in (9). This is the general case in which the attenuation approaches zero with increasing frequency.

In this, as in the special cases considered, the equations reveal that the anomalous falling off of the attenuation constant with frequency may be ascribed to the indefinite increase of one of the transverse dimensions of the guide, with the consequent disappearance of all terms that normally account for the rise with frequency. Thus all surfaces at which lines of electric force begin or end are relegated to regions infinitely far apart from one another where their existence can no longer be of any practical consequence. In the rectangular case, the guide ceases thereby to be of finite height, but in the two circular cases, where the field is wrapped around an axis or around a circular cylinder, the guide retains a finite form and the field within it, so constrained, continues to display the properties characteristic of a field of infinite range.

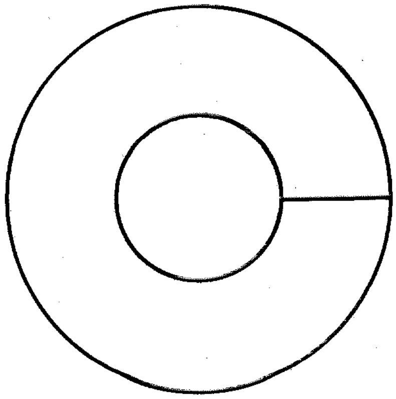

Fig. 12—Guide composed of two coaxial circular cylinders and a radial baffle plane.

Exact Design and Analysis of Double- and Triple-Tuned Band-Pass Amplifiers*

By MILTON DISHAL

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THE PURPOSE of this paper is to present a quick, complete, and exact method of design and analysis of double- and triple-tuned band-pass amplifiers.

The necessary small-percentage pass-band equations are derived giving the relationship between the circuit characteristics and the response characteristics. These circuit characteristics are: the resonant frequency f_0, coefficient of coupling K, the circuit Q, and the input and output capacitances C_{in} and C_{out}. The response characteristics are: the percentage bandwidth between peaks $\Delta f_p/f_0$, the peak-to-valley response ratio within the pass band V_p/V_v, the peak-to-"skirt" response ratio V_p/V at different skirt-to-peak bandwidth ratio points $\Delta f/\Delta f_p$ outside the pass band, the circuit gain at the peaks, and the phase shift θ at any frequency.

These design equations, extended to the case of one to eight cascaded stages, are incorporated in two sets of conveniently used nomographs, one set for the double-tuned circuits and one set for triple-tuned circuits. Specific examples of the use of these nomographs are given.

• • •

1. Symbols

f_0 = resonant frequency (see Section 5)
ω_0 = resonant radian frequency
n = decrement of a resonant circuit (see Section 4)
Q = reciprocal of decrement (see Section 4)
K_C = coefficient of capacitive coupling between resonant circuits (see Fig. 3)
K_L = coefficient of inductive coupling between resonant circuits (see Fig. 3)
K_M = coefficient of mutual inductive coupling between resonant circuits (see Fig. 3)
$K = [K_c(\omega/\omega_0) - K_L(\omega_0/\omega)]$
C, L, M, R = capacitance (farads), inductance (henries), mutual inductance (henries), resistance (ohms)
B = susceptance
g = conductance
$F = (f/f_0 - f_0/f) = (\omega/\omega_0 - \omega_0/\omega)$
θ = phase angle between a resulting voltage and the driving current, or the phase angle between a resulting current and the driving voltage
Δf = difference between two frequencies
V = response voltage
β = a constant for double-tuned circuits (a function of peak-to-valley ratio)
N = number of cascaded stages
γ = a constant for triple-tuned circuits (a function of peak-to-valley ratio)
Δf_p = bandwidth between response peaks
V_p = voltage at peaks of the response
V_v = voltage at the valley of the response
$D = n_2/n_1 = Q_1/Q_2$

2. Introduction

To aid in the design and analysis of circuits which produce a band-pass response with respect to frequency, there has arisen a large body of literature[1–4] under the two general headings of filter theory and coupled-circuit theory.

However, it would be worthwhile to have collected in one place a method of design that will quickly and easily give answers to questions of the following type which might arise in the course of a thorough design of, say, a wide-band intermediate-frequency amplifier for receivers:

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* Reprinted from *Proceedings of the I.R.E.*, v. 35, pp. 606–626; June, 1947.

[1] E. S. Purington, "Single and Coupled Circuit Systems," *Proceedings of the I.R.E.*, v. 18, pp. 983–1016; June, 1930.

[2] C. B. Aiken, "Two Mesh Tuned Coupled Circuit Filters," *Proceedings of the I.R.E.*, v. 25, pp. 230–272; February, 1937.

[3] F. X. Rettenmeyer, "Radio Bibliography—Filters," *Radio*, n. 273, pp. 26–30; October, 1942.

[4] T. E. Shea, "Transmission Networks and Wave Filters," D. Van Nostrand Company, Inc., New York, N. Y., 1929.

A. For a given bandwidth and "flatness" of response, exactly how much more gain can be obtained if we use triple-tuned rather than double-tuned circuits?

B. Can a certain skirt-selectivity specification be satisfied using only five double-tuned circuits? If so, what must the circuit constants be? What peak-to-valley ratio will there be in the pass band?

C. How much more gain and how much greater skirt selectivity will be obtained if we accept a relatively poor response in the pass band by allowing a 1.3 peak-to-valley ratio in preference to a good 1.05 peak-to-valley ratio? What must the circuit constants be for both cases?

D. Will more gain per stage be obtained if all the loading is done in one of the resonant circuits, or should the Q of all the resonant circuits be made equal?

In this paper, through the medium of two sets of three nomographs each, the writer hopes to provide in one place a ready means of obtaining exact answers (with a minimum of time and calculation) to the above and other questions for the case of double-tuned and triple-tuned band-pass circuits when small-percentage (20 percent or less) bandwidths are used.

The concepts and constants used are those commonly associated with coupled-circuit theory. Filter-theory constants and concepts are always useful, and when many tuned circuits are coupled together it is practically necessary to use the filter-theory type of design. However, for both double- and triple-tuned circuits, it is possible to obtain exact closed-form solutions for the circuit response (when band-pass percentages are approximately 20 percent or less); and these solutions are more concisely stated in terms of coupled-circuit constants.

The circuit constants used are the resonant frequency f_0, the Q of each resonant circuit used, and the coefficient of coupling K between resonant circuits. The response constants are the percentage bandwidth between peaks $\Delta f_p/f_0$, the peak-to-valley ratio V_p/V_v inside the pass band, (this fixes the goodness or "flatness" of the pass band); and the skirt bandwidths $\Delta f/\Delta f_p$ at different skirt response points V_p/V (this fixes the sharpness of cutoff or the skirt selectivity outside the pass band), the circuit gain at the peaks, and the phase shift at any frequency.

The results of the double-tuned analysis (i.e., the nomographs and the family of phase-shift curves) will be given next with examples of their use.

3. *Design and Analysis of Double-Tuned Circuits by Means of the Nomographs*

From (19a), A_1, C_1, D_2, and E_1 of this paper, a set of nomographs have been prepared, and a family of curves have been prepared from the phase-shift equation (26). The use of these nomographs is best explained by a few specific examples.

3.1 EXAMPLE I

Knowing that the gain per stage is approximately

$$\text{Gain} = g_m/4\pi\Delta f_p\sqrt{C_1C_2}$$

and that $C_1 \doteq C_2 \doteq 10$ micromicrofarads and $g_m = 5\times10^{-3}$ mho, it is decided that five stages are probably needed to obtain a certain desired gain. A ratio of peak gain to valley gain of 1.10 will be satisfactory. A bandwidth between peaks Δf_p of 2 megacycles is required; and to make the percentage bandwidth approximately 20 percent or less, a midfrequency f_0 of 30 megacycles is chosen.

What loading resistances should be used to give the proper Q in the two tuned circuits? What exact gain per stage will be obtained? What must the mutual impedance be to give the proper coefficient of coupling? What will the bandwidth be 6 decibels down from the peaks? What will the bandwidth be 60 decibels down from the peaks?

Starting with Chart A, place a straight edge between point 5 on the "Number of Cascaded Stages (N)" column and point 1.10 on the "(V_p/V_v)" column. From the "[$Q/(f_0/\Delta f_p)$]" column, we find that the Q of each resonant circuit must be

$$Q = 0.69\frac{f_0}{\Delta f_p} \doteq 10,$$

and from this same column the gain per stage will be

$$\text{Gain} = 0.69\frac{g_m}{4\pi\Delta f_p\sqrt{C_1C_2}} \doteq 14.$$

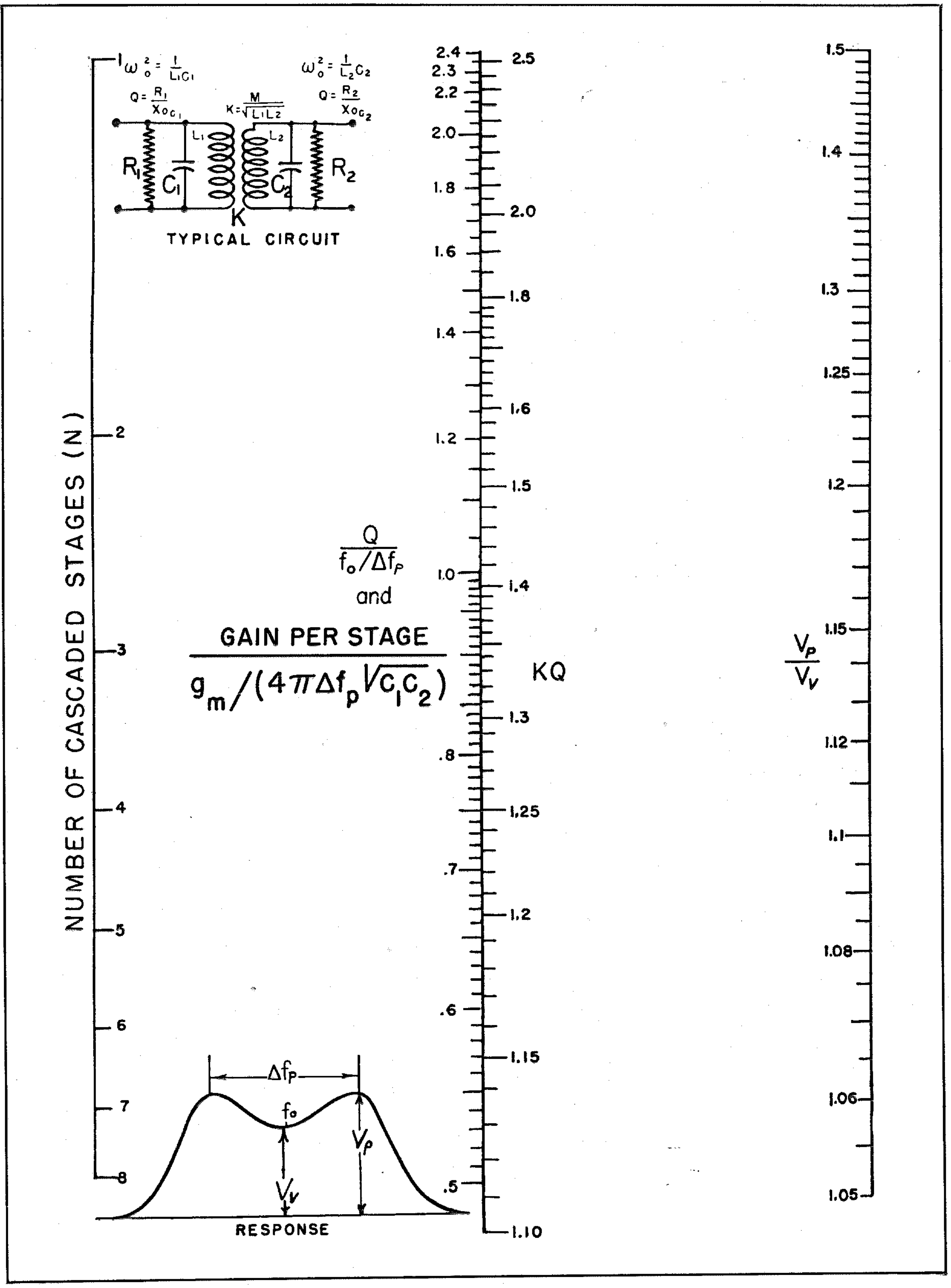

Chart A—Double-tuned band-pass circuit design.

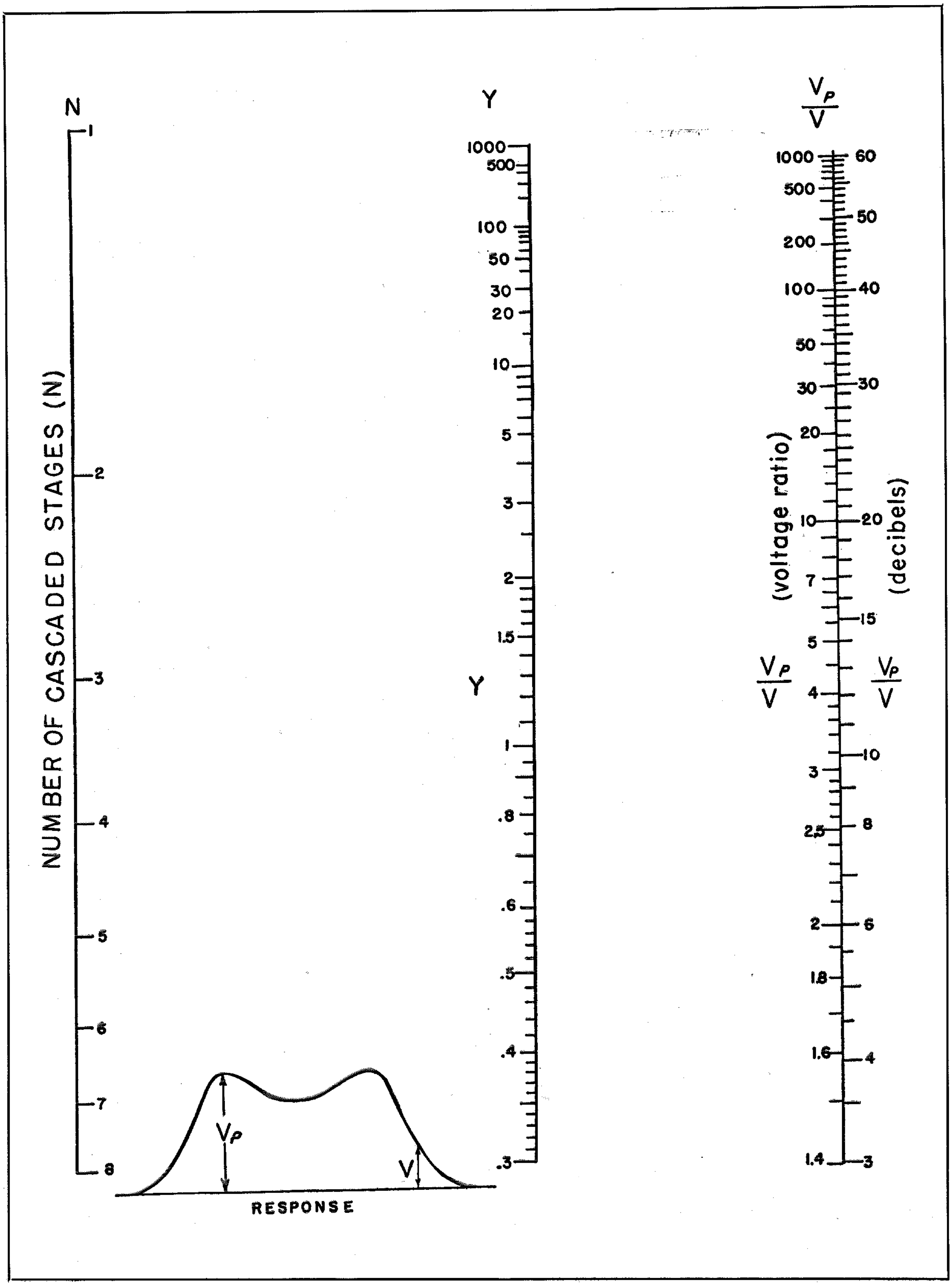

Chart B—Double-tuned band-pass circuit design.

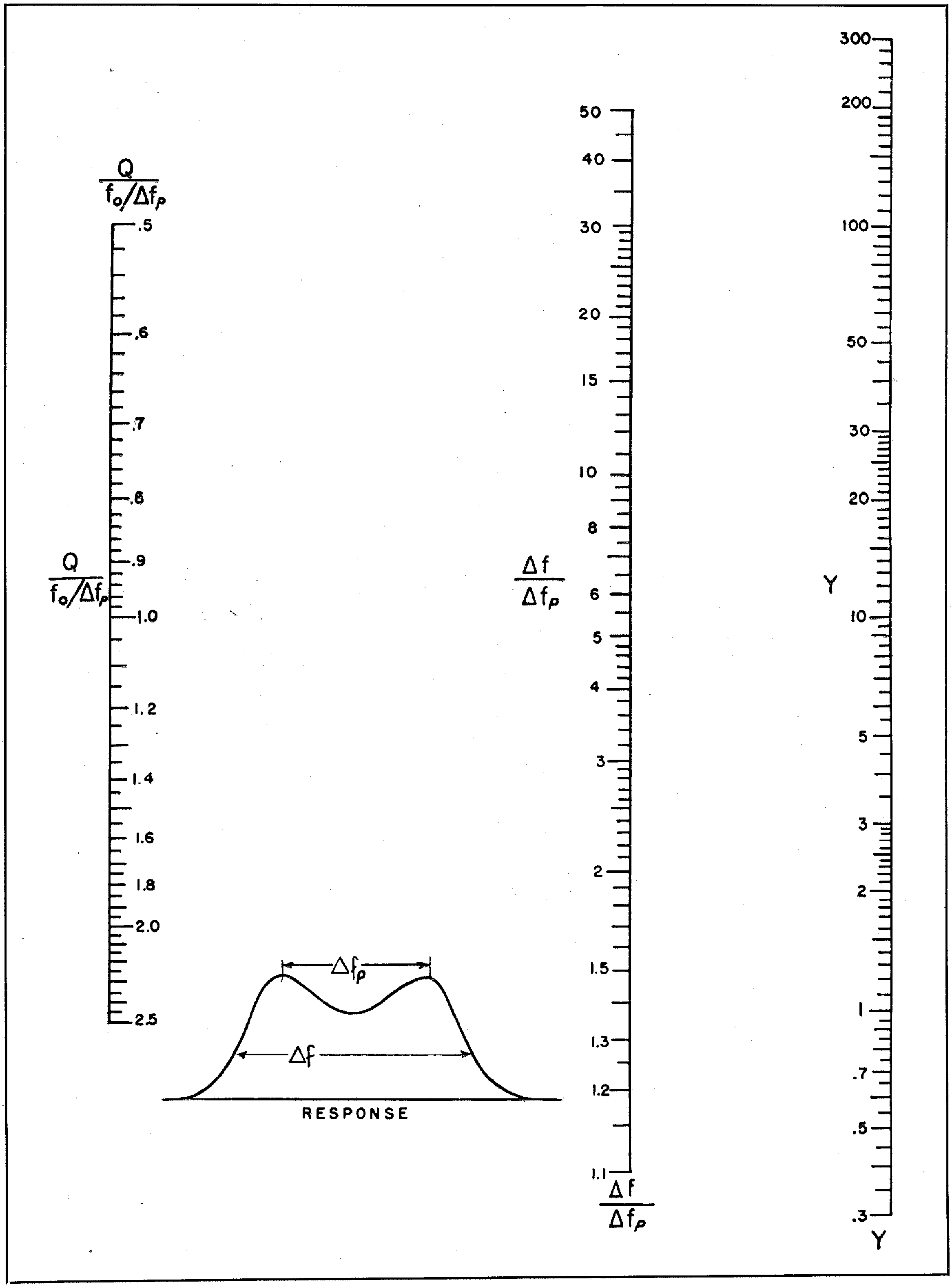

Chart C—Double-tuned band-pass circuit design.

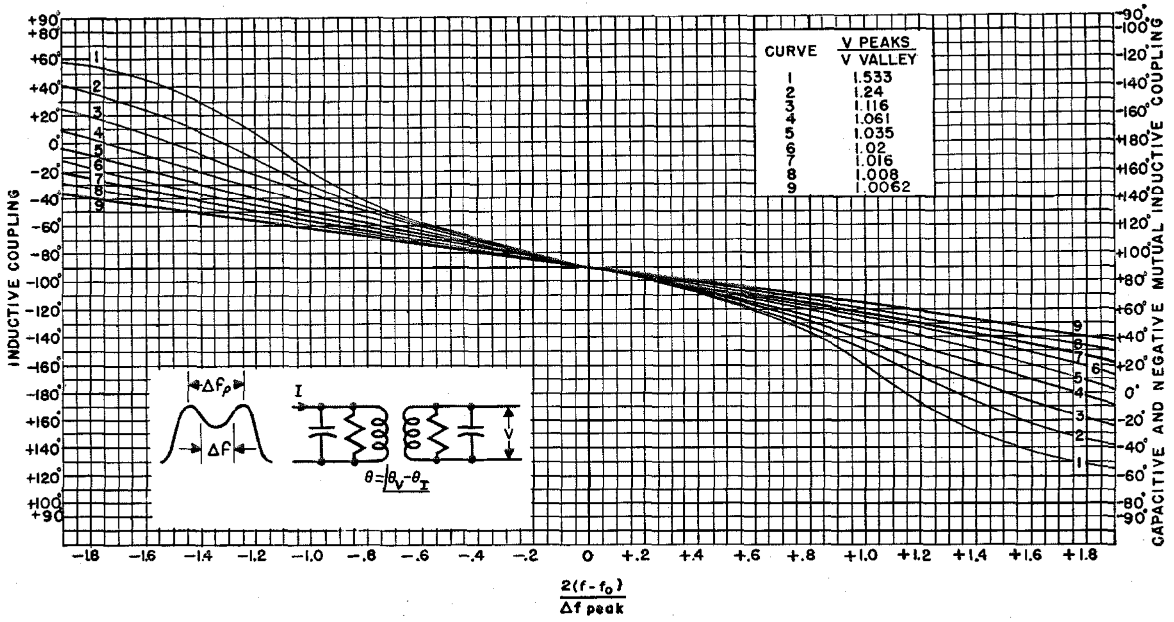

Chart D—Phase shift for a flat-top double-tuned circuit for different peak-to-valley ratios.

Knowing the necessary resonant-circuit Q and the reactance of the total shunt capacitances in the resonant circuits, the necessary resultant loading resistance is, of course, given simply by $R = QX_{0_c} = 10 \times 500 = 5000$ ohms.

Since the coils used will usually have appreciable loss, they will effectively supply a shunt loading resistance of value $Q_L X_{0_c}$ where Q_L is the Q of the inductance and X_{0_c} is the impedance of the shunt capacitance at the resonant frequency.

Thus, the resistance R_+ which must be added in parallel with the above effective resistance, due to a Q_L of 50, for example, to produce the required resultant Q is

$$R_+ = \left[Q \Big/ \left(1 - \frac{Q}{Q_L}\right)\right] X_{0_c} \approx 6250 \text{ ohms.}$$

From the "KQ" column of Chart A, the coefficient of coupling must be

$$K = \frac{1.22}{Q} = 0.122.$$

In the type of circuit chosen (see Figs. 1 and 2), the mutual reactance between the two resonant circuits is then found from the simple equation for the coefficient of coupling as given with each type of coupling in Fig. 3.

To consider skirt selectivity, use Charts B and C. On Chart B, place a straight edge between point 5 on the "Number of Cascaded Stages (N)" column, and 6 decibels (or 2) on the "(V_p/V)" column. Read 0.56 on the middle, or "Y" column. Now, going to Chart C, place the straight

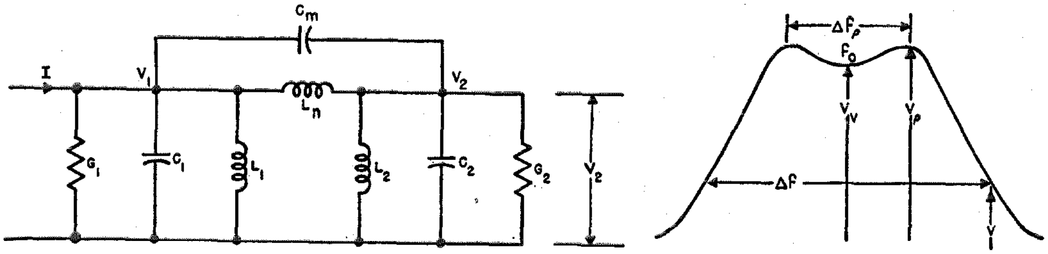

Fig. 1—Basic double-tuned two-node band-pass circuit using both inductive and capacitive coupling and the type of voltage response to be considered.

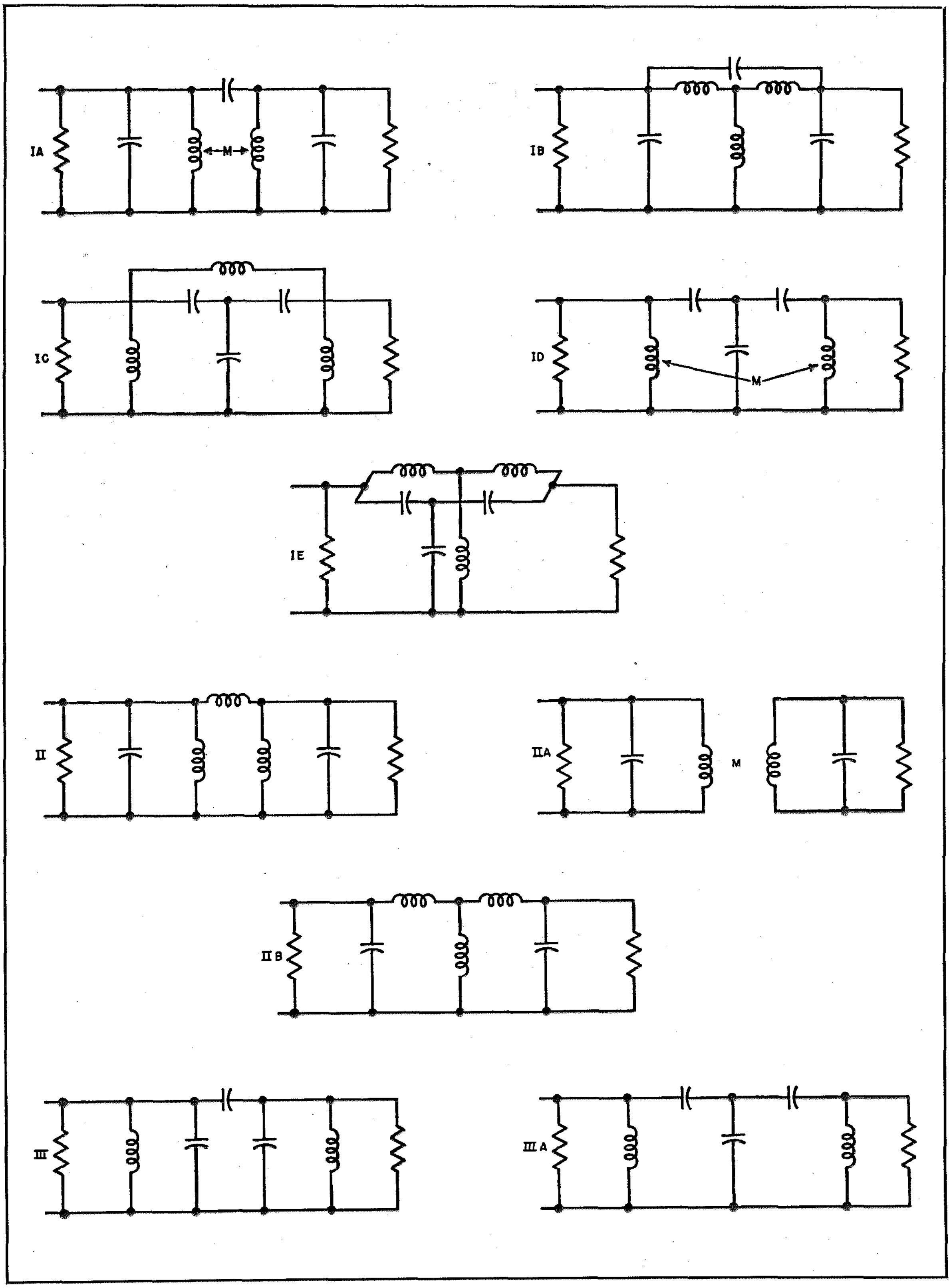

Fig. 2—Ten two-node circuits. The circuit of Fig. 1 is exactly equivalent to these circuits.

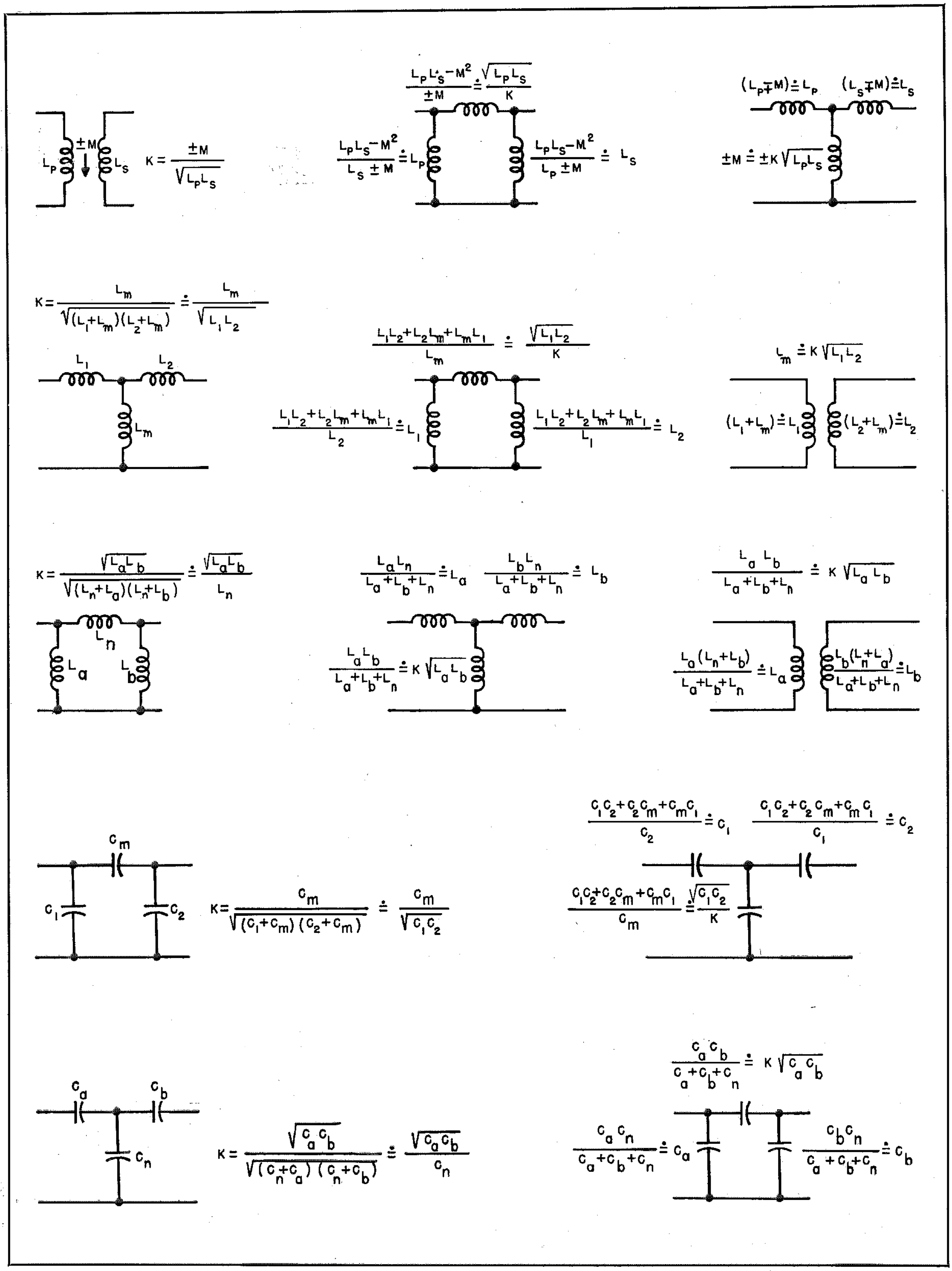
L_p
$\pm M$
L_s
$K = \frac{\pm M}{\sqrt{L_pL_s}}$
$\frac{L_pL_s - M^2}{\pm M} \doteq \frac{\sqrt{L_pL_s}}{K}$
$\frac{L_pL_s - M^2}{L_s \pm M} \doteq L_p$
$\frac{L_pL_s - M^2}{L_p \pm M} \doteq L_s$
$(L_p \mp M) \doteq L_p$
$(L_s \mp M) \doteq L_s$
$\pm M \doteq \pm K\sqrt{L_pL_s}$
$K = \frac{L_m}{\sqrt{(L_1+L_m)(L_2+L_m)}} \doteq \frac{L_m}{\sqrt{L_1L_2}}$
L_1
L_2
L_m
$\frac{L_1L_2+L_2L_m+L_mL_1}{L_m} \doteq \frac{\sqrt{L_1L_2}}{K}$
$\frac{L_1L_2+L_2L_m+L_mL_1}{L_2} \doteq L_1$
$\frac{L_1L_2+L_2L_m+L_mL_1}{L_1} \doteq L_2$
$L_m \doteq K\sqrt{L_1L_2}$
$(L_1+L_m) \doteq L_1$
$(L_2+L_m) \doteq L_2$
$K = \frac{\sqrt{L_aL_b}}{\sqrt{(L_n+L_a)(L_n+L_b)}} \doteq \frac{\sqrt{L_aL_b}}{L_n}$
L_n
L_a
L_b
$\frac{L_aL_n}{L_a+L_b+L_n} \doteq L_a$
$\frac{L_bL_n}{L_a+L_b+L_n} \doteq L_b$
$\frac{L_aL_b}{L_a+L_b+L_n} \doteq K\sqrt{L_aL_b}$
$\frac{L_aL_b}{L_a+L_b+L_n} \doteq K\sqrt{L_aL_b}$
$\frac{L_a(L_n+L_b)}{L_a+L_b+L_n} \doteq L_a$
$\frac{L_b(L_n+L_a)}{L_a+L_b+L_n} \doteq L_b$
C_m
C_1
C_2
$K = \frac{C_m}{\sqrt{(C_1+C_m)(C_2+C_m)}} \doteq \frac{C_m}{\sqrt{C_1C_2}}$
$\frac{C_1C_2+C_2C_m+C_mC_1}{C_2} \doteq C_1$
$\frac{C_1C_2+C_2C_m+C_mC_1}{C_1} \doteq C_2$
$\frac{C_1C_2+C_2C_m+C_mC_1}{C_m} \doteq \frac{\sqrt{C_1C_2}}{K}$
C_a
C_b
C_n
$K = \frac{\sqrt{C_aC_b}}{\sqrt{(C_n+C_a)(C_n+C_b)}} \doteq \frac{\sqrt{C_aC_b}}{C_n}$
$\frac{C_aC_b}{C_a+C_b+C_n} \doteq K\sqrt{C_aC_b}$
$\frac{C_aC_n}{C_a+C_b+C_n} \doteq C_a$
$\frac{C_bC_n}{C_a+C_b+C_n} \doteq C_b$

Fig. 3—Coefficient of coupling used in the analysis and the π, T, and transformer exact equivalents, and approximations for small couplings.

edge between 0.69 on the "$[Q/(f_0/\Delta f_p)]$" column and 0.56 on the "Y" column and read from the middle column that

$$\Delta f_{6\text{ decibels}} = 1.95\Delta f_p = 3.9 \text{ megacycles.}$$

The bandwidth at the 60-decibels-down point is obtained in exactly the same way, i.e., on Chart B, place the straight edge between the point 5 on the "Number of Cascaded Stages (N)" column and 60 decibels (or 1000) on the "(V_p/V)" column. Read 3.6 on the "Y" column. Going to Chart C, place the straight edge between point 0.69 on the "$[Q/(f_0/\Delta f_p)]$" column and 3.6 on the "Y" column. Read from the "$[(\Delta f/\Delta f_p)]$" column that

$$\Delta f_{60\text{ decibels}} = 4.4\Delta f_p = 8.8 \text{ megacycles.}$$

Any other points on the response curve are found in the same manner.

3.2 EXAMPLE II

Knowing that the approximate gain per stage is Gain$=g_m/4\pi\Delta f_p\sqrt{C_1C_2}$, it is decided that only 3 stages are needed to give a certain desired gain. It is necessary that the skirt selectivity be such that the bandwidth 60 decibels down be only 5 times the bandwidth between the peaks, i.e., $\Delta f_{60\text{ decibels}}/\Delta f_p = 5$. What must be the Q of each tuned circuit to obtain this skirt selectivity? What exact gain per stage will be obtained? What coefficient of coupling is required? What peak-to-valley ratio must be accepted to obtain this selectivity?

Starting with Chart B, place a straight edge between point 3 in the "Number of Cascaded Stages (N)" column and point 60 decibels on the "(V_p/V)" column and read 9.6 from the "Y" column. Going to Chart C, place the straight edge between point 5 on the "$(\Delta f/\Delta f_p)$" column and 9.6 on the "Y" column and read on the "$[Q/(f_0/\Delta f_p)]$" column that the required Q is

$$Q = 1.1\frac{f_0}{\Delta f_p}.$$

Now going to Chart A, place the straight edge between point 3 on the "Number of Cascaded Stages (N)" column and 1.1 on the "$[Q/(f_0/\Delta f_p)]$" column. The exact gain will be

$$\text{Gain} = 1.1\frac{g_m}{4\pi\Delta f_p\sqrt{C_1C_2}}$$

and, from the "KQ" column, the required coefficient of coupling is

$$K = \frac{1.48}{Q}.$$

From the "(V_p/V_v)" column, the resulting peak-to-valley ratio will be

$$V_p/V_v = 1.27.$$

3.3 EXAMPLE III

The nomographs may be conveniently used for analysis of coupled circuits, as well as for design or synthesis.

Thus, given the Q of two resonant circuits as 20, the coefficient of coupling K between them as 0.085, and the resonant frequency as 15 megacycles, what is the response curve?

The product of KQ is 1.7. Going to Chart A, set the straight edge between 1 on the "Number of Cascaded Stages (N)" column and 1.7 on the "KQ" column. From the "(V_p/V_v)" column, $V_p/V_v = 1.15$. From the "$[Q/(f_0/\Delta f_p)]$" column, the bandwidth between peaks will be

$$\Delta f_p = 1.38\frac{f_0}{Q} = 1.04 \text{ megacycles.}$$

To find the width of the skirts at different points, e.g., 10 times or 20 decibels down, go to Chart B. Place the straight edge between 1 on the "Number of Cascaded Stages (N)" column and 20 decibels on the "(V_p/V)" column and read 10 on the "Y" column. Going to Chart C, place the straight edge between 1.38 on the "$[Q/(f_0/\Delta f_p)]$" column and 10 on the "Y" column and see that

$$\Delta f_{20\text{ decibels}} = 4.4\Delta f_p = 4.6 \text{ megacycles.}$$

Any other points on the skirts are obtained in the same way.

3.4 EXAMPLE IV

To find the phase shift at any point in the pass band, Chart D is used.

It should be noted that $2(f-f_0)/\Delta f_p$ (which is the abscissa of the graph) is merely a way of writing $(\Delta f/\Delta f_p)$ to show more clearly that in the phase-shift equation (26), Δf defines *two* frequencies equidistant from the resonant frequency. The abscissa is $(+)$ for frequencies above the resonant frequency and is $(-)$ for frequencies below the resonant frequency. For example, at the high-frequency peak, $f=f_{ph}$ and $2(f-f_0)/\Delta f_p=+1$ and at the low-frequency peak, $f=f_{pl}$ and $2(f-f_0)/\Delta f_p=-1$.

Note also that the ordinates give the phase shift *per stage*. If N cascaded identical stages are used, this phase shift is then multiplied by N.

Finally, note that the peak-to-valley ratios for each curve are the ratios for a single *stage*.

Thus, if 6 cascaded stages are being used to produce a resultant peak-to-valley ratio of 1.05, each single stage must have a peak-to-valley ratio equal to the 6th root of 1.05, or 1.0083. For this case, Curve 8 would therefore give the phase shift versus frequency per stage. This phase shift at each frequency is then multiplied by 6 to give the resultant phase-shift-versus-frequency curve.

4. Circuits Which Are Analyzed

The basic circuit analyzed is the two-node network consisting of two resonant circuits coupled together both inductively and capacitively. This circuit and the response investigated are shown in Fig. 1.

By virtue of the exact equivalence of π's, T's, and transformers, the exact analysis of the basic circuit is immediately applicable to ten more circuits. These ten circuits are shown in Fig. 2 and the equations giving the values of the equivalent elements are given in Fig. 3. Lattice, bridged-T, etc., equivalents may also be used.

By virtue of the concept of duality,[5] the analysis of the basic two-node network is immediately applicable to the dual two-mesh network given in Fig. 4, where I, the equivalent constant-current generator, and g, C, and L, are substituted for by e, R, L, and C, respectively. Again, by virtue of the equivalence of π's, T's, and transformers, the analysis also applies to the ten additional circuits given in Fig. 5. Thus, a total of 22 band-pass circuits are effectively analyzed in this paper, plus any lattice, bridged-T, etc., equivalents which the reader may desire to use.

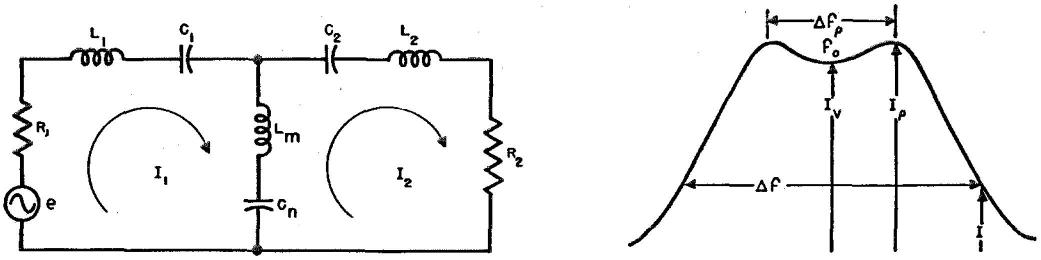

Fig. 4—Basic double-tuned two-mesh band-pass circuit (using both inductive and capacitive coupling) and the type of current response to be considered.

The two-node circuit of Fig. 1 is picked as the circuit to be analyzed, rather than the dual two-mesh circuit of Fig. 4, because vacuum-tube amplifiers are effectively high-impedance generators, and for practically all high-frequency band-pass amplifier applications, high-impedance resonance is desired as obtained by the use of the circuits of Figs. 1 and 2.

If very-small-percentage pass bands are to be produced, and very slight inequality in the height of the two peaks can be tolerated, then all 22 of the circuits shown in Figs. 1, 2, 4, and 5 can be used as either high- or low-impedance circuits by means of the following reasoning. For the small-percentage band-pass case, it is convenient (and correct) to consider the band-pass characteristic as being produced in the following manner:

A. Fundamentally, the configuration of only the lossless reactive components produces the band-pass response; the percentage bandwidth being fixed (to a first approximation) by the coefficient of coupling K. Figs. 1 and 2 and Figs. 4 and 5 give the two-node and the two-mesh reactive networks that can produce a band-pass characteristic. (Consider the shunt resistors of Figs. 1 and 2 to be open-circuited and the series resistors of Figs. 4 and 5 to be short-circuited.)

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[5] Electrical Engineering Staff, Massachusetts Institute of Technology, "Electric Circuits," John Wiley and Sons, New York, N. Y., 1940, pp. 245–246.

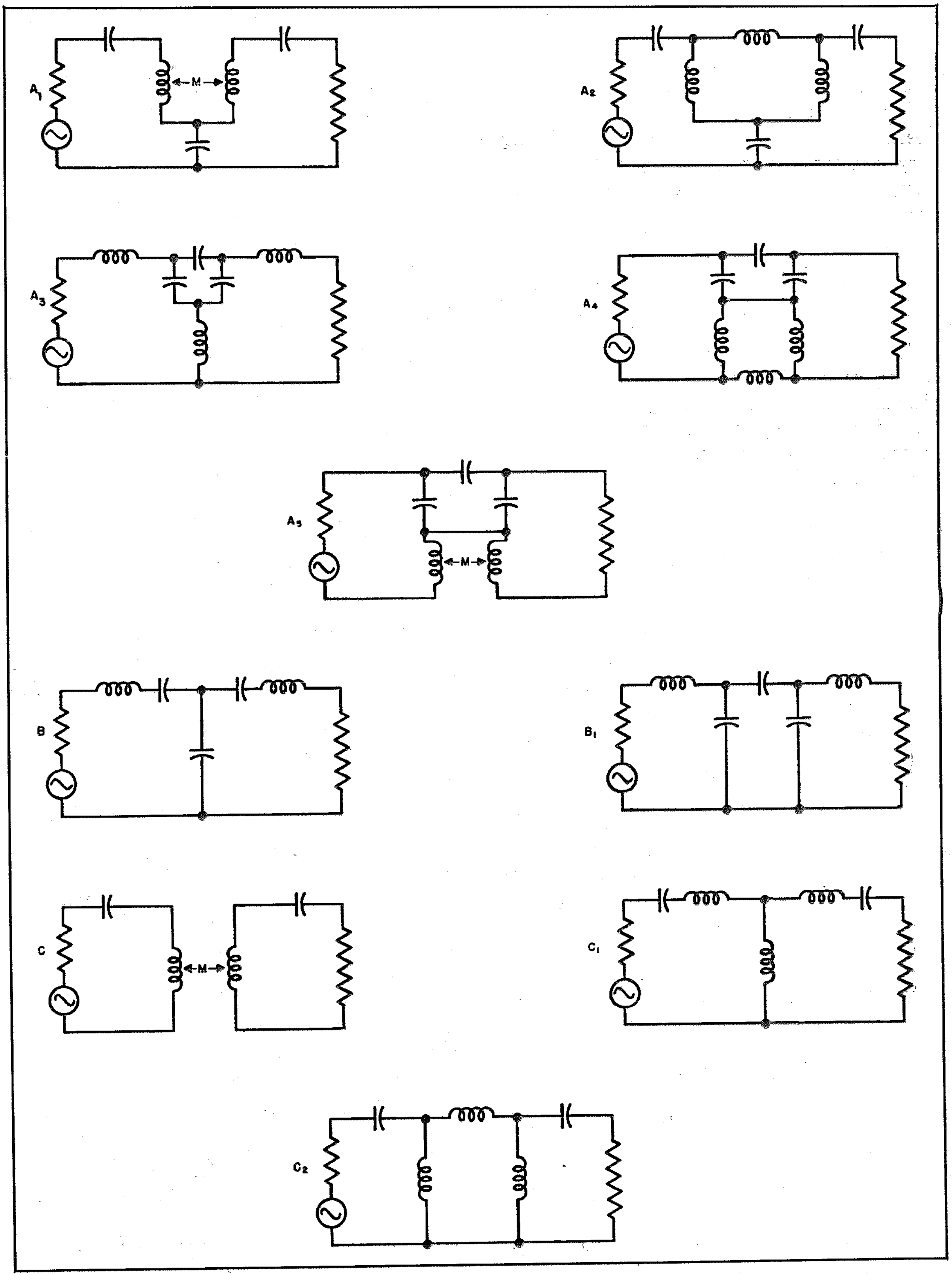

Fig. 5—Ten two-mesh circuits. The circuit of Fig. 4 is exactly equivalent to these circuits.

B. The peak-to-valley ratio is fixed to a first approximation by the required Q of the input and output resonant circuits. The correct resonant circuit Q can be produced in three ways: (1) by placing a small resistance in series with the resonant circuit, $Q=X_{0_c}/R_s$; (2) by placing a large resistance in parallel with the resonant circuit, $Q=R_p/X_{0_c}$, or (3) by a combination of both series and parallel loading. For this case

$$Q=\frac{1}{(R_s/X_{0_c})+(X_{0_c}/R_p)}.$$

C. The driving force may be applied in two ways: either an infinite-impedance (i.e., zero conductance) constant-current generator may be placed in parallel with either the resonating inductance or the resonating capacitance (never across the mutual reactance); or a zero-impedance constant-voltage generator may be placed in series with either the resonating inductance or resonating capacitance (never in series with the mutual reactance).

In practice, all equivalent generators have finite output impedances associated with them. Thus, the above steps, B and C, are interrelated to the extent that the effect of the output impedance of the generator upon the resonant circuit Q must be considered.

D. The output voltage may be obtained across either the resonating inductance or the resonating capacitance in the output circuit. Of course, we must consider the effect of the resistive component of the load upon the Q of the output resonant circuit.

5. *Elements Which Are Resonated*

It is important to know exactly what elements are resonated in the above circuits. The elements which are tuned to resonance in circuit 1, Fig. 5, and all the two-node circuits are indicated by the following procedure: node 2 is shorted to ground and all the reactive elements remaining are resonated at the desired frequency; then node 1 is shorted to ground (the short on node 2 is removed) and all the remaining reactive elements are resonated to the above frequency. Thus, in circuit 1, C_1 plus C_m is resonated with the resultant of L_2 and L_n in parallel. Thus, for Fig. 1, we have

$$\omega_0^2=\frac{1}{\left(\frac{L_1L_n}{L_1+L_n}\right)(C_1+C_m)}=\frac{1}{\left(\frac{L_2L_n}{L_2+L_n}\right)(C_2+C_m)}. \qquad (1)$$

This method of defining the resonances also introduces a very practical method of aligning double- or triple-tuned coupled resonant circuits. *First*, completely detune all but one of the resonant circuits without affecting the mutual impedance. This detuning effectively short-circuits the node to ground for all practical purposes, and may be accomplished simply by placing an additional capacitance across the resonant circuits whose value is approximately three or four times that of the capacitance in the circuit. Or, if iron-slug tuning is used, sufficient detuning can usually be accomplished merely by turning the slug to its extreme position. *Second*, feed a signal into the circuit at the desired resonant frequency and tune the remaining circuit, which is not detuned, for maximum output. This procedure is then repeated until all the circuits have been resonated in the above manner.

Actually, for a certain distribution of the circuit constants, i.e., $Q_1=Q_2$, there is a more convenient method of alignment which will be mentioned later.

In the dual two-mesh circuits, the elements to be resonated are indicated by the following procedure: mesh 2 is open-circuited and all the reactances remaining in the circuit are resonated. Then, with mesh 2 returned to its normal condition, mesh 1 is open-circuited and all remaining reactive elements are resonated to the same frequency. Thus, for Fig. 4, we have:

$$\omega_0^2=\frac{1}{(L_1+L_m)\left(\frac{C_1C_n}{C_1+C_n}\right)}=\frac{1}{(L_2+L_m)\left(\frac{C_2C_n}{C_2+C_n}\right)}. \qquad (2)$$

6. *Exact Response Equations*

The node equations for the circuit shown in Fig. 1 are:

$$\left.\begin{aligned} I&=[g_1+j(B_{c_1}+B_{c_m}-B_{L_1}-B_{L_n})]V_1\\ &\quad -j(B_{c_m}-B_{L_n})V_2\\ 0&=-j(B_{c_m}-B_{L_n})V_1\\ &\quad +[g_2+j(B_{c_2}+B_{c_m}-B_{L_2}-B_{L_n})]V_2.\end{aligned}\right\} \tag{3}$$

As mentioned in Section 4, the solution of the above two equations for the response voltage V_2 contains the solution for all 22 circuits shown in Figs. 1, 2, 4, and 5.

A great simplification is produced in the resulting equations for the circuits if the resonant frequency f_0, the coefficient of coupling K between resonant circuits, and the decrement of each resonant circuit n are introduced into the circuit equations. (The decrement is the reciprocal of the more commonly used Q.)

With the introduction of these constants, the equations can be expressed in terms of the three quantities only instead of in terms of the eight L, C, and R elements making up the circuit. Our mental picture of the circuit action is thus greatly simplified.

By solving (3) for the output voltage V_2 and introducing into the solution the three constants mentioned above, namely,

$$\begin{aligned}\omega_0^2&=\frac{1}{\left(\dfrac{L_1L_n}{L_1+L_n}\right)(C_1+C_m)}\\ &=\frac{1}{\left(\dfrac{L_2L_n}{L_2+L_n}\right)(C_2+C_m)}\end{aligned} \tag{4}$$

$$n_1=\frac{g_1}{\omega_0(C_1+C_m)} \tag{5}$$

$$n_2=\frac{g_2}{\omega_0(C_2+C_m)} \tag{6}$$

$$K_C=\frac{C_m}{\sqrt{(C_1+C_m)(C_2+C_m)}} \tag{7}$$

$$K_L=\frac{\sqrt{L_1L_2}}{\sqrt{(L_1+L_n)(L_2+L_n)}}, \tag{8}$$

we obtain as the exact solution for the magnitude of the response

$$\begin{aligned}V_2&=\frac{I}{\omega_0\sqrt{(C_1+C_m)(C_2+C_m)}}\\ &\quad\times\frac{K}{\sqrt{F^4-2\left[K^2-\dfrac{n_1^2+n_2^2}{2}\right]F^2+(K^2+n_1n_2)^2}}\end{aligned} \tag{9}$$

and the phase of the output voltage with respect to the constant current source is

$$\tan\theta=\frac{\pm[K^2+n_1n_2-F^2]}{\pm[(n_1+n_2)F]} \tag{10}$$

where

$$K=\left(K_C\frac{\omega}{\omega_0}-K_L\frac{\omega_0}{\omega}\right) \tag{11}$$

$$F=\left(\frac{\omega}{\omega_0}-\frac{\omega_0}{\omega}\right). \tag{12}$$

The sign to be used in the phase-shift equation (10), is the sign of the quantity

$$\left(K_C\frac{\omega}{\omega_0}-K_L\frac{\omega_0}{\omega}\right).$$

Thus, with capacitive coupling predominant, the top signs are used in numerator and denominator, and, with positive inductive coupling predominant, the bottom signs are used.

Examination of the numerator of (9) shows immediately one characteristic of the response. The numerator becomes zero and thus there is a null response at:

$$\frac{\omega_{\text{null}}}{\omega_0}=\sqrt{\frac{K_L}{K_C}}. \tag{13}$$

With reference to circuit IC of Fig. 2, it should be mentioned that if the winding sense of the inductances is such that the mutual inductive coupling "aids" the capacitive coupling there is no null of response, for then the sign of K_L in (11) is negative (−) and, therefore, the numerator never becomes zero.

We will now introduce into the above exact equations the approximations that produce the symmetrical and relatively simple small-percentage pass-band analysis.

7. *Small-Percentage Pass-Band Response Shape*

Because K in (9) is a function of frequency, the exact response shape is not symmetrical either geometrically or arithmetically with respect to frequency. If, however, we limit ourselves to small-percentage pass bands where ω/ω_0 varies in value over the small range from, say, 0.9 to 1.1 then two important simplifications immediately result in the factors shown in (11) and (12).

Equation (11) becomes independent of frequency:

$$K \doteq (K_C - K_L). \tag{11a}$$

(It must be realized that this approximation cannot be used in the region of the null given by (13).)

Equation (12) becomes

$$F = \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) = \frac{(\omega+\omega_0)}{2\omega} \times \frac{2(\omega-\omega_0)}{\omega_0} \doteq \frac{2(\omega-\omega_0)}{\omega_0} \doteq \frac{\Delta f}{f_0}, \tag{12a}$$

where Δf is the frequency bandwidth between points equidistant from the resonant frequency f_0.

With the above limitation, (9) shows that in the small-percentage pass-band case (where (11a) applies) the shape of the amplitude response curve is independent of the type of coupling used. The gain obtained with inductive coupling only is slightly greater than that obtained with capacitive coupling, for, as seen from (9), the capacitances that must be considered in figuring the gain are (C_1+C_m) and (C_2+C_m). C_m is the equivalent high-side capacitances of Fig. 5 and is zero for inductive coupling only.

The phase shift as given by (10) does differ for the two types of coupling. Since the top signs are used with capacitive coupling and the bottom signs with inductive coupling, we will have positive phase angles with capacitive coupling and negative phase angles with inductive coupling.

The frequency at which the response maximum and minimum occurs is given by differentiating (9) with respect to F (i.e., $\Delta f/f_0$) and equating to zero. This results in

$$\left(\frac{\Delta f \text{ peak}}{f_0}\right)^2 = K^2 - \frac{n_1^2+n_2^2}{2}, \tag{14}$$

and the location of the minimum is given by $\Delta f_v/f_0 = 0$.

The response at the peaks, obtained by substituting (14) in (9), is

$$V_{\text{peaks}} = \frac{I}{\omega_0\sqrt{(C_1+C_m)(C_2+C_m)}} \times \frac{K}{\sqrt{K^2(n_1+n_2)^2 + n_1^2 n_2^2 - \left(\frac{n_1^2+n_2^2}{2}\right)^2}}. \tag{15}$$

The response at the minimum or valley, which is at the resonant frequency, is obtained from (9) by setting $\Delta f/f_0 = 0$ and is

$$V_v = \frac{I}{\omega_0\sqrt{(C_1+C_m)(C_2+C_m)}} \frac{K}{K^2+n_1 n_2}. \tag{16}$$

The peak-to-valley ratio is, therefore,

$$\left(\frac{V_p}{V_v}\right) = \frac{K^2+n_1 n_2}{\sqrt{K^2(n_1+n_2)^2 + n_1^2 n_2^2 - \left(\frac{n_1^2+n_2^2}{2}\right)^2}}. \tag{17}$$

What we desire, so far as design is concerned, is the values of the decrement n (or Q) and the coefficient of coupling K required to give a certain peak-to-valley ratio. By combining (14) and (17), we obtain

$$\frac{n_1+n_2}{2} = \beta\left(\frac{\Delta f_p}{f_0}\right), \tag{18}$$

where

$$\beta = \sqrt{\frac{1}{2}\left[\frac{(V_p/V_v)_1}{\sqrt{(V_p/V_v)_1^2 - 1}} - 1\right]}, \tag{19}$$

and where the subscript 1 is to show that this is the peak-to-valley ratio of one double-tuned stage. Equation (18) is one of the desired design equations and shows that the required average of the decrements of the primary and secondary is fixed only by the peak-to-valley ratio desired and the percentage bandwidth.

The smaller we desire the peak-to-valley ratio to be (thus the flatter the response is in the pass band) the larger β becomes and, therefore, the greater must be the average decrement, i.e., the lower must be the Q. From (19), we find that β varies between the values of 1.75 to 0.42 as the peak-to-valley ratio varies, respectively, between the values of 1.01 to 1.50.

Now, making use of equations (14) and (18), we obtain for the required coefficient of coupling

$$K=\frac{\Delta f_p}{f_0}\sqrt{1+\beta^2\frac{2(1+D^2)}{(1+D)^2}} \qquad (20)$$

where β is given by (22) and D is the ratio of the primary Q to the secondary Q.

$$D=\frac{n_2}{n_1}=\frac{Q_1}{Q_2}.$$

Thus, we see that the coefficient of coupling required is fixed mainly by the percentage band pass desired and is also dependent (not to a great extent, however) on the ratio D of primary Q to secondary Q. Equation (20) is the second of our desired design equations.

Dividing equation (20) by equation (18), we obtain

$$\frac{K}{(n_1+n_2)/2}=\sqrt{\frac{2(1+D^2)}{(1+D)^2}+\frac{1}{\beta^2}}. \qquad (21)$$

This is a very useful equation because it does not involve frequency. It shows that as soon as the peak-to-valley ratio (i.e., β) and the Q ratio are fixed, then the ratio of the coefficient of coupling K and the average decrement $(n_1+n_2)/2$ is also fixed, and conversely for a given circuit where the Q ratio and the ratio of the coefficient of coupling and the average decrement is fixed, the peak-to-valley ratio is fixed. It should be understood that the Q ratio D has an almost second-order effect; for the quantity $2(1+D^2)/(1+D)^2$ is equal to unity when the Q ratio is unity, and approaches a maximum value of two when the Q ratio approaches either zero or infinity.

The next design equation desired is one that will give the output voltage or the gain of the circuit at the peaks of the response. By substituting the design conditions given by (18) and (20) in the equation giving the response at the peaks, which is (15), we obtain

$$V_p=\frac{1}{\beta}\times\frac{I}{4\pi\Delta f_p\sqrt{(C_1+C_m)(C_2+C_m)}}\times\sqrt{\frac{1+\frac{2(1+D^2)}{(1+D)^2}\beta^2}{1+\beta^2}} \qquad (22)$$

and for the usual case, where the constant-current generator of value I is a vacuum tube, $I=g_mE_g$, and we have

$$\text{Gain}_{\text{(per stage)}}=\frac{1}{\beta}\times\frac{g_m}{4\pi\Delta f_p\sqrt{(C_1+C_m)(C_2+C_m)}}\times\sqrt{\frac{1+\frac{2(1+D^2)}{(1+D)^2}\beta^2}{1+\beta^2}}. \qquad (23)$$

Design equation (23) brings out several points of interest with reference to the gain obtained with "flat-topped" band-pass circuits. We see that the gain depends directly on the g_m of the tube used and inversely on the *numerical* bandwidth desired between peaks Δf_p. The midfrequency has no effect on the gain (as long as the bandwidth Δf_p is a small percentage of the midfrequency f_0). The gain is also inversely proportional to the square root of the product of the total capacitance across the input or output circuits that must be resonated. We see also that the gain is inversely proportional to the factor β which is given by (19) and which is a measure of the flatness of response in the flat-top pass band. The flatter the pass band, the lower the gain obtainable. Finally, the gain depends on the square root of a quantity involving the ratio of primary Q to secondary Q.

This square root has only an almost second-order effect on the gain. It is interesting, however, to see the effect of this Q ratio on the gain. If Q_1 equals Q_2, the factor under discussion becomes unity. If Q_2 is made infinite and all the loading is done on the primary side, we obtain

$$\sqrt{\frac{1+2\beta^2}{1+\beta^2}}$$

and, if Q_1 is made infinite and all the loading is done on the secondary side, we again obtain

$$\sqrt{\frac{1+2\beta^2}{1+\beta^2}}.$$

In most practical designs, β will have a value close to unity; therefore, if all loading is done on one side of the band-pass circuit, approximately 25 percent more gain per stage will be obtained, as compared to the case where the primary and secondary are equally loaded (i.e., $Q_1=Q_2$).

It may be mentioned here that practical considerations dealing with ease of circuit alignment, and "Miller effect" detuning, lead to the conclusion that in many cases it is better to make $Q_1 = Q_2$ and thus sacrifice the above 25 percent additional gain per stage. These points will be discussed later.

The next desired design equation is concerned with the shape of the circuit response outside the pass band, i.e., the skirt selectivity. By combining (9), giving the response at any frequency, and (15), giving the response at the peaks, and (18) and (20), giving the required circuit constants, we obtain for the ratio of peak response V_p to the response V, at any bandwidth Δf,

$$\left(\frac{V_p}{V}\right)_1 = \sqrt{1 + \left[\frac{(\Delta f/\Delta f_p)^2 - 1}{2\beta\sqrt{1+\beta^2}}\right]^2} \tag{24}$$

and solving (24) for $\Delta f/\Delta f_p$, we obtain

$$\frac{\Delta f}{\Delta f_p} = \sqrt{1 \pm 2\beta\sqrt{1+\beta^2}\sqrt{(V_p/V)_1^2 - 1}}, \tag{25}$$

where the subscript 1 is to show that the ratios are the voltage ratios for *one* double-tuned stage.

This is the last of our desired design equations and we see that the larger β is made (therefore, the flatter the response inside the pass band) the wider are the skirts at any skirt-response point, i.e., skirt selectivity becomes poorer as the pass-band response is improved. It should be noted from (24) or (25) that, for a given peak-to-valley ratio (i.e., a given β), the shape of the response curve is independent of the ratio of primary Q to secondary Q.

The plus-or-minus sign in (25) should also be noted. When the plus sign is used, we obtain the skirt bandwidths outside the response peaks, and when the minus sign is used, we obtain the bandwidths inside the peaks of the response curve.

To make analysis as complete as possible, the phase of the response voltage with respect to the driving current should also be given. By combining (10) for the phase shift with design equations (18) and (20), we obtain

$$\tan \theta_{\text{per stage}} = \frac{\pm[1 + 2\beta^2 - (\Delta f/\Delta f_p)^2]}{\pm[2\beta(\pm\Delta f/\Delta f_p)]}. \tag{26}$$

In (26) the top sign is used in front of the numerator and denominator when $(K_C - K_L)$ is plus, i.e., with a net capacitive coupling. (It should be remembered that these equations should not be applied to the region in the vicinity of the null given by (13).) The plus sign is used inside the bracket in the denominator for the frequencies above the resonant frequency and the minus sign is used for the frequencies below the resonant frequency.

From (26) we can see that for *inductive* coupling the phase shift at the midfrequency (i.e., $\Delta f = 0$) is -90 degrees; at the low-frequency peak, the tangent of the phase angle is $(-/+)\beta$; and at the high-frequency peak, the tangent of the phase angle is $(-/-)\beta$. Since in many applications satisfactory flatness in the pass band is given when β is approximately unity, we see that the phase shift at the low-frequency peak is usually approximately -45 degrees and the high-frequency peak usually has a phase angle of approximately -135 degrees.

With *capacitive* coupling, we see that the phase shift at the midfrequency is $+90$ degrees; the tangent of the phase angle at the low-frequency peak is $(+/-)\beta$; the tangent of the phase angle at the high-frequency peak is $(+/+)\beta$; and for β equal approximately to unity the phase shift at the low-frequency peak is thus approximately $+135$ degrees, and at the high-frequency peak it is approximately $+45$ degrees.

It should be noted that for a given peak-to-valley ratio (i.e., a given β), the phase shift is independent of the Q ratio.

8. Small-Percentage Pass-Band Design Equations When $Q_1 = Q_2$

Design equations having even a small degree of complexity are, in many cases, not used by engineers. However, conveniently used graphical representations of the complex equations will usually be put to use.

Usually, identical band-pass circuits are cascaded to produce intermediate-frequency-amplifier chains. Various applications may necessitate the use of from one to, perhaps, eight cascaded stages. It would appear worthwhile to develop an exact, rapid, graphical method of designing cascaded circuits so that they produce a specified response shape.

Since the number of cascaded stages used must be one of the design parameters, consideration of (18), (20), (21), (22), (25), and (26) shows that some form of family-of-curves representation or its equivalent is necessary. We further note that (20) and (22) are complicated by the relatively second-order effect of the Q ratio which would necessitate an almost useless family of curves. Because of this complication, we will consider the case where $Q_1=Q_2$, in the graphical method of design; and the equations themselves can be used directly when Q_1 does not equal Q_2.

There are two important practical reasons why a design using $Q_1=Q_2$ should be used whenever possible. The first reason is concerned with the problem of aligning cascaded flat-topped band-pass circuits. The second reason is concerned with the detuning effect caused by the fact that the input and output capacitances of a pentode change with gain-control setting due to plate-to-grid capacitance feedback (Miller effect) and space-charge effects.

With reference to the alignment of cascaded flat-topped circuits, if Q_1 is made equal to Q_2 the circuits can be aligned just as single-peaked or single-tuned circuits are aligned, i.e., by using a single-frequency signal generator (not a "sweeper"), and tuning for absolute maximum output. With double-peaked circuits, the signal generator is set at the frequency at which the *low* peak of the response is desired and all the circuits are tuned *lower* in frequency for maximum response. (Or the signal generator may be set at the frequency at which it is desired to have the *high*-frequency peak, and all the circuits are then tuned *higher* in frequency for maximum response.) It can be shown that if Q_1 equals Q_2, equal absolute maxima of response are obtained at the peaks only when both circuits are tuned to the same resonant frequency (as described in Section 4) and, conversely, when both circuits are tuned to the same resonant frequency, absolute maximum (and equal) response is obtained at both peaks (so long as there is no loss in the mutual reactance). This fact is the basis of the method of alignment just described.

When Q_1 does not equal Q_2, tuning of the circuits to produce an absolute maximum of response at one frequency would necessitate the two circuits being tuned to different resonant frequencies and the two peaks are then of different amplitudes.

With reference to the second reason for making Q_1 equal to Q_2, it is desirable to have a response curve which is not affected when the gain (i.e., the g_m) of the amplifier tubes is changed. Unfortunately, the changes in input and output capacitances of a pentode, with changing g_m (due to plate-to-grid capacitance feedback and space-charge effects), detunes the resonant circuits. However, it can be shown that, with $Q_1=Q_2$, a slight detuning of the resonant circuits will have practically negligible effect on the *symmetry* of the response curve. Thus, although the response curve as a whole will move slightly as the gain control is changed, the shape of the curve will remain sensibly constant when $Q_1=Q_2$.

When circuits are cascaded, the voltage responses at a given frequency are multiplied together to give the resultant voltage response. When the cascaded circuits are all identical, it is obvious that to obtain the resultant voltage response, the voltage response of one circuit is raised to that power given by the number of cascaded circuits.

We must realize that all the voltage responses in the previous equations apply to only one double-tuned stage. If we are going to cascade N stages and want a certain resultant peak-to-valley ratio V_p/V_v, the peak-to-valley ratio of each circuit $(V_p/V_v)_1$, must equal $(V_p/V_v)^{1/N}$. Likewise, if the resultant skirt-response ratio for N cascaded stages is to be V_p/V, then the skirt-response ratio for each stage $(V_p/V)_1$, must equal $(V_p/V)^{1/N}$.

Thus, in (19) and (25), which apply to one stage only, we should make the above substitutions to make them apply to N cascaded stages.

For the case of $Q_1=Q_2$, the design equations then become as follows:

Let

$$\beta=\sqrt{\frac{1}{2}\left[\frac{(V_p/V_v)^{1/N}}{\sqrt{(V_p/V_v)^{2/N}-1}}-1\right]}\cdot \tag{19a}$$

Then

$$\frac{Q}{f_0/\Delta f_p}=\frac{1}{\beta} \tag{18a}$$

$$KQ=\sqrt{1+1/\beta^2} \tag{21a}$$

$$\left(\frac{V_p}{V}\right)=\left\{1+\left[\frac{(\Delta f/\Delta f_p)^2-1}{2\beta\sqrt{1+\beta^2}}\right]^2\right\}^{N/2} \tag{24}$$

$$\frac{\text{Gain per stage}}{g_m/4\pi\Delta f_p\sqrt{(C_1+C_m)(C_2+C_m)}}=\frac{1}{\beta} \tag{23a}$$

$$\frac{\Delta f}{\Delta f_p}=\sqrt{1\pm 2\beta\sqrt{1+\beta^2}\sqrt{(V_p/V)^{2/N}-1}} \tag{25a}$$

$$\tan\theta_{\text{per stage}}=\frac{\pm[1+2\beta^2-(\Delta f/\Delta f_p)^2]}{\pm[2\beta(\pm\Delta f/\Delta f_p)]}. \tag{26}$$

9. *Formation of Triple-Tuned Band-Pass Circuits*[6,7]

Any two of the circuit configurations shown in Figs. 1 and 2 may be connected in series to form a triple-tuned three-node band-pass circuit. (This also means, of course, that one of the circuits shown can be used twice.) Similarly, the circuit configurations of Figs. 4 and 5 can be used to form three-mesh band-pass circuits.

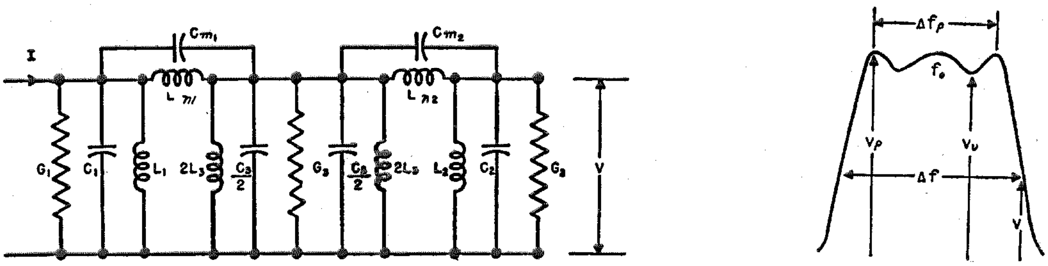

Fig. 6—Basic triple-tuned three-node band-pass circuit using both inductive and capacitive coupling and the type of voltage response to be considered.

With respect to the calculation of the two equal coefficients of coupling which appear in the resulting triple-tuned circuit, maximum gain will be obtained if the following procedure is used: the middle resonant circuit formed when two of the node networks of Figs. 1 and 2 are connected in series should be considered to be formed from two identical resonant circuits in parallel (i.e., each one having twice the net inductance and one half the net capacitance). The input resonant circuit is then coupled to one of the above resonant circuits and the output circuit is coupled to the other resonant circuit.

The middle resonant circuit formed when two of the mesh circuits of Figs. 4 and 5 are connected in series should be considered to be formed from two identical resonant circuits in series (i.e., each one having twice the net capacitance and half the next inductance). The input resonant circuit is then coupled to one of the above resonant circuits and the output circuit is coupled to the other resonant circuit.

The points made in Sections 4 and 5 of the double-tuned analysis apply also to the triple-tuned case, and, rather than repeat them here, it will be assumed that the reader will again refer to the above sections.

To obtain a flat-topped response with three peaks of equal amplitude in the pass band, all the loading must be removed from the middle tuned circuit, which is formed when two double-tuned circuits are thus connected in series. Otherwise, as will be shown later, the outer two peaks of the response will be lower in amplitude than the middle peak.

Unfortunately, it is often impossible to obtain inductances of sufficient Q for the middle tuned circuit unless extremely large coil forms and shield cans are used. It will be shown that the required Q for the input and output tuned circuits is of the order of the value of the reciprocal of the percentage bandwidth. Thus, if a bandwidth between peaks of 400 kilocycles is desired with a midfrequency of 20 megacycles, the required Q of the input and output circuit will be approximately 50. To approach the ideal triple-tuned response curve, the Q of the middle tuned circuit must be of the order of 10 times (or more) the Q of the input and output circuits. Thus, a Q of the order of 500 or more is required in the above case. It is difficult to obtain an inductance of this Q.

However, if the midfrequency of the 400-kilocycle pass band were shifted down to 4 megacycles, the required Q of the input and output circuits would then be about 10, and the necessary middle-circuit Q would be at least 100. This Q can be obtained without too much trouble. Thus, if triple-tuned band-pass circuits are to be

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[6] E. A. Guillemin, "Communication Networks," John Wiley and Sons, New York, N. Y., v. 1, 1931; pp. 335–339. This analysis deals with the rather unfortunate case (in so far as good band-pass response is concerned) of $Q_1 = Q_2 = Q_3$.

[7] M. R. Winkler, "A 3 Resonant Circuit Transformer," *Electronics*, v. 16, pp. 96–100; January, 1943. Here again the main emphasis is placed on the case of $Q_1 = Q_2 = Q_3$.

used, it would be worthwhile choosing a 10-percent, or even greater, bandwidth.

As in the double-tuned case, the high-impedance or node circuits will be considered to be used the most, and therefore the specific analysis will be made using three node circuits having both inductive and capacitive coupling, as shown in Fig. 6. It should be clearly realized, however, that the resulting analysis applies exactly to all of the myriad triple-tuned networks that can be formed from the networks of Figs. 1 and 2 and 4 and 5.

10. Exact Triple-Tuned Response Equation

The node equations which apply to the triple-tuned circuit of Fig. 6 are:

$$I=\left\{g_1+j\left[\omega(C_1+C_{m_1})-\frac{1}{\omega\left(\dfrac{L_1L_n}{L_1+L_{n_1}}\right)}\right]\right\}V_1-j\left(\omega C_{m_1}-\frac{1}{\omega L_{n_1}}\right)V_2+0$$

$$0=-j\left(\omega C_{m_1}-\frac{1}{\omega L_{n_1}}\right)V_1+\left\{g_3+j\left[\omega(C_3+C_{m_1}+C_{m_2})-\frac{1}{\omega\left(\dfrac{L_{n_1}L_3L_{n_2}}{L_n L_3+L_3L_{n_2}+L_{n_1}L_{n_2}}\right)}\right]\right\}V_2-j\left(\omega C_{m_2}-\frac{1}{\omega L_n}\right)V_3$$

$$0=0-j\left(\omega C_{m_2}-\frac{1}{\omega L_{n_2}}\right)V_2+\left\{g_2+j\left[\omega(C_2+C_{m_2})-\frac{1}{\omega\left(\dfrac{L_2L_{n_2}}{L_2+L_{n_2}}\right)}\right]\right\}V_3. \tag{27}$$

Introducing the resonant frequency (as defined in Section 5), the coefficient of coupling, and the decrement, we obtain from (27) the complete, exact solution for the magnitude and phase of the output voltage.

$$\frac{V_3}{I/\omega_0\sqrt{(C_1+C_{m_1})(C_2+C_{m_2})}}$$

$$=\frac{\frac{1}{2}K^2}{\left\{F^6-[2K^2-(n_1^2+n_2^2+n_3^2)]F^4 +\{K^4-K^2[n_1^2+n_3^2-n_2(n_1+n_3)]+n_1^2n_2^2+n_2^2n_3^2+n_3^2n_1^2\}F^2+\left[K^2\left(\dfrac{n_1+n_3}{2}\right)+n_1n_2n_3\right]^2\right\}^{\frac{1}{2}}} \tag{28}$$

$$\tan\theta=\frac{+F[K^2+(n_1n_2+n_2n_3+n_3n_1)-F^2]}{-\left[K^2\left(\dfrac{n_1+n_3}{2}\right)+n_1n_2n_3-(n_1+n_2+n_3)F^2\right]} \tag{29}$$

where as before $F=\left(\dfrac{\omega}{\omega_0}-\dfrac{\omega_0}{\omega}\right)$ $K=\left(K_C\dfrac{\omega}{\omega_0}-K_L\dfrac{\omega_0}{\omega}\right)$.

11. Small-Percentage Band-Pass Design Equations

Applying the reasoning used in Section 6 of the double-tuned analysis, we will consider the small-percentage band-pass case, i.e., where ω/ω_0 becomes only about 10 percent greater or less than unity. We thus have the two great simplifications:

$$K\doteq(K_C-K_L) \quad\text{and}\quad \left(\frac{\omega}{\omega_0}-\frac{\omega_0}{\omega}\right)\doteq\frac{\Delta f}{f_0}.$$

Setting the derivative with respect to

$$F\left(\doteq\frac{\Delta f}{f_0}\right)$$

of (28) equal to zero, we obtain for the location of the maxima (plus sign) and minima (minus sign)

$$\frac{\Delta f_{\mathrm{max}}}{f_0}=0. \qquad \frac{\Delta f_{\mathrm{min\ max}}}{f_0}=\frac{2}{3}\left[K^2\frac{n_1^2+n_2^2+n_3^2}{2}\right.$$

$$\left.\pm\tfrac{1}{3}\{K^4-K^2[n_1^2+n_3^2+n_2(3n_1+4n_2+3n_3)]+(n_1^4+n_2^4+n_3^4)-(n_1^2n_2^2+n_2^2n_3^2+n_3^3n_1^2)\}^{\frac{1}{2}}\right]. \tag{30}$$

We will obtain the design equations for the case where the Q's of the input and output circuits are the same ($n_1=n_3=n$) and the middle resonant circuit Q is much greater than the Q of the input and output circuits ($n_2 \ll n$).

For this case, the general response (28) becomes the relatively simple equation

$$\frac{V_3}{I/\omega_0[(C_1+C_{m_1})(C_2+C_{m_2})]^{\frac{1}{2}}}=\frac{\frac{1}{2}K^2}{\left[\left(\frac{\Delta f}{f_0}\right)^6-2(K^2-n^2)\left(\frac{\Delta f}{f_0}\right)^4+(K^2-n^2)^2\left(\frac{\Delta f}{f_0}\right)^2+K^4n^2\right]^{\frac{1}{2}}} \tag{28a}$$

and from (30) the locations of the maxima and minima are given by

$$\left(\frac{\Delta f_{\mathrm{max}}}{f_0}\right)^2=0 \quad \text{and} \quad (K^2-n^2) \tag{30a}$$

$$\left(\frac{\Delta f_{\mathrm{min}}}{f_0}\right)^2=\tfrac{1}{3}(K^2-n^2)=\tfrac{1}{3}\left(\frac{\Delta f_{\mathrm{max}}}{f_0}\right)^2$$

and the phase-shift equation becomes

$$\tan\theta=\frac{+\left(\frac{\pm\Delta f}{f_0}\right)\left[K^2+n^2-\left(\frac{\Delta f}{f_0}\right)^2\right]}{-\left[K^2n-2n\left(\frac{\Delta f}{f_0}\right)^2\right]}. \tag{29a}$$

Substituting the locations of the maxima (30a) into the response equation (28a), gives the response at the peaks, which is

$$V_{\mathrm{peaks}}=\frac{I}{\omega_0[(C_1+C_{m_1})(C_2+C_{m_2})]^{\frac{1}{2}}}\times\frac{1}{2n}. \tag{31}$$

Substituting the location of the minimum (30a) in the response equation (28a) gives the following response at the valley:

$$V_{\mathrm{valley}}=\frac{I}{\omega_0[(C_1+C_{m_1})(C_2+C_{m_2})]^{\frac{1}{2}}}$$

$$\times\frac{\frac{1}{2}K^2}{[4/27(K^2-n^2)^3+K^4n^2]^{\frac{1}{2}}} \tag{32}$$

and so the peak-to-valley ratio is

$$\left(\frac{V_p}{V_v}\right)_1=\frac{[4/27(K^2-n^2)^3+K^4n^2]^{\frac{1}{2}}}{K^2n}. \tag{33}$$

Introducing the location of the outside peaks (30a) into the peak-to-valley ratio (33), we can solve for the decrement that is required to produce a desired peak-to-valley ratio with a given percentage bandwidth between outside peaks. The result is our first design equation.

$$\frac{Q}{f_0/\Delta f_p}=\frac{1}{\gamma} \tag{34}$$

where

$$\gamma=\frac{\left\{\frac{1+(V_p/V_v)_1}{[(V_p/V_v)_1^2-1]^{\frac{1}{2}}}\right\}^{\frac{1}{3}}+\left\{\frac{1-(V_p/V_v)_1}{[(V_p/V_v)_1^2-1]^{\frac{1}{2}}}\right\}^{\frac{1}{3}}}{3^{\frac{1}{2}}} \tag{35}$$

and, using (30a), we have as the equation giving the coefficient of coupling which is required to obtain a given peak-to-valley ratio with a given percentage bandwidth between peaks

$$\frac{K}{\Delta f_p/f_0}=(1+\gamma^2)^{\frac{1}{2}} \tag{36}$$

where γ is given by (35).

Multiplying (34) by (36), we obtain our second design equation.

$$KQ=[1+(1/\gamma^2)]^{\frac{1}{2}}. \tag{37}$$

The next desired design equation is the one giving the skirt selectivity. Substituting the design equations (34) and (36) into the response equation (28a), we obtain the response at any point in terms of the peak-to-valley ratio (represented by γ of (27)) and the percentage bandwidth at the outside peaks. Dividing the result

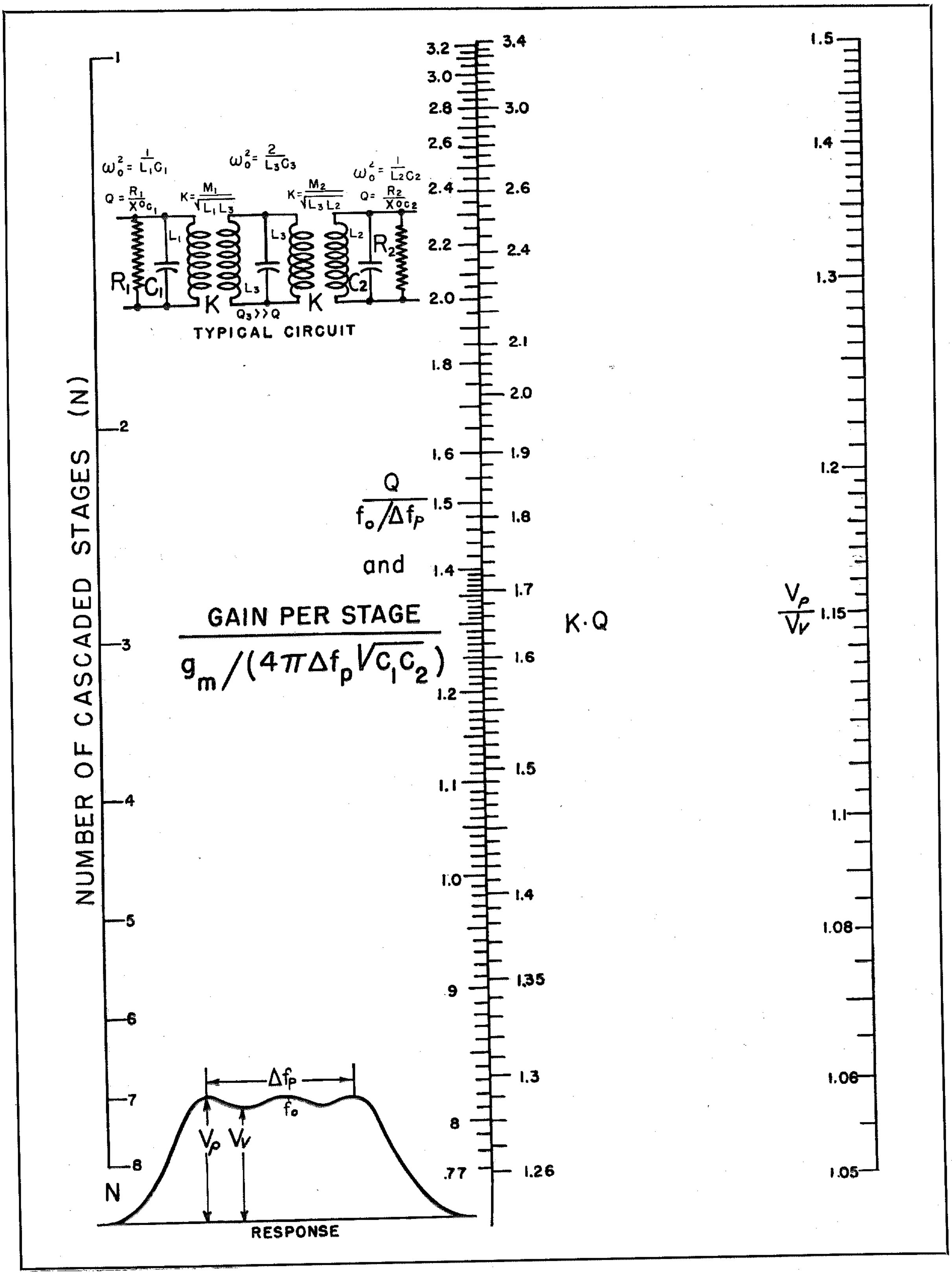

Chart I—Triple-tuned band-pass circuit design.

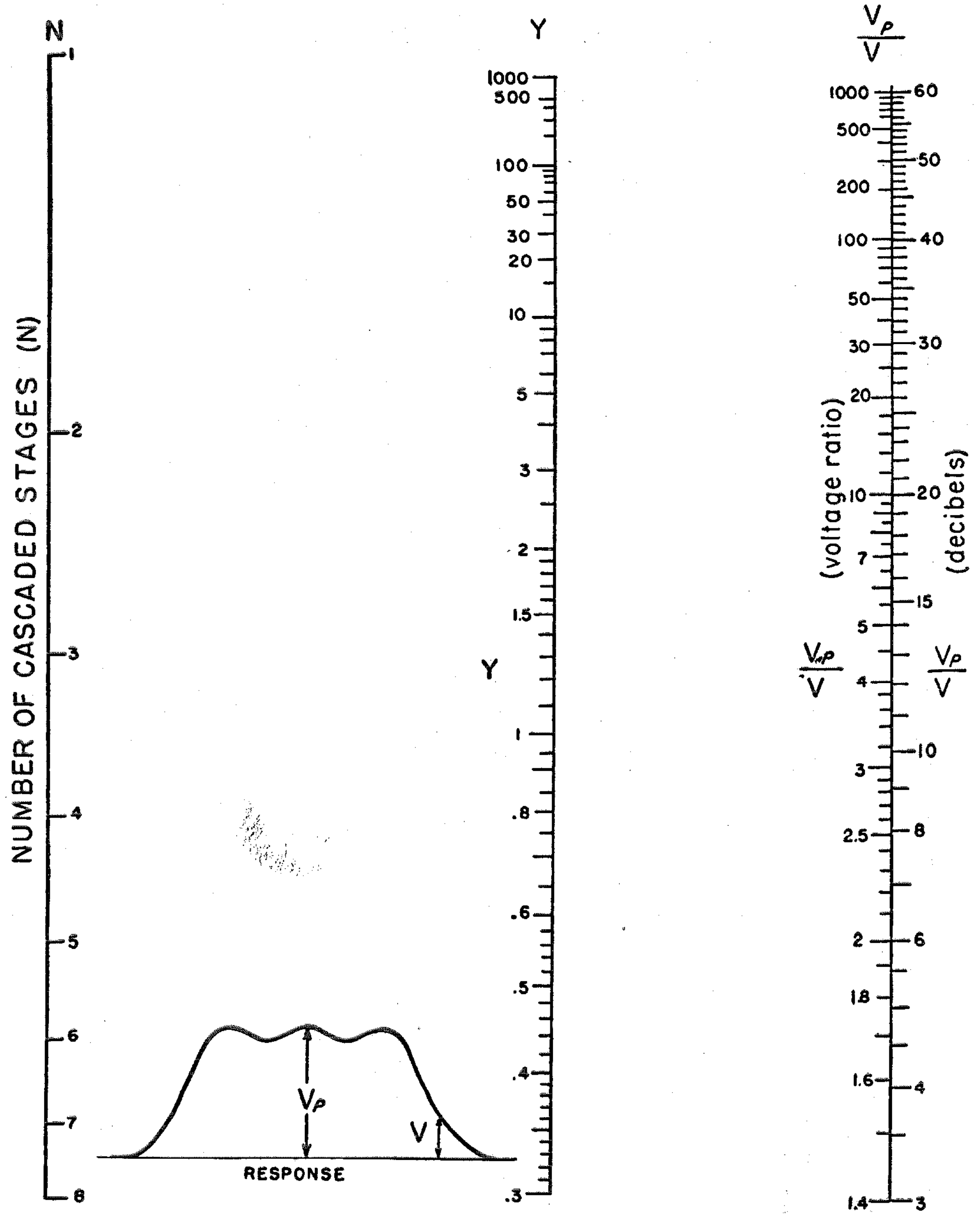

Chart II—Triple-tuned band-pass circuit design.

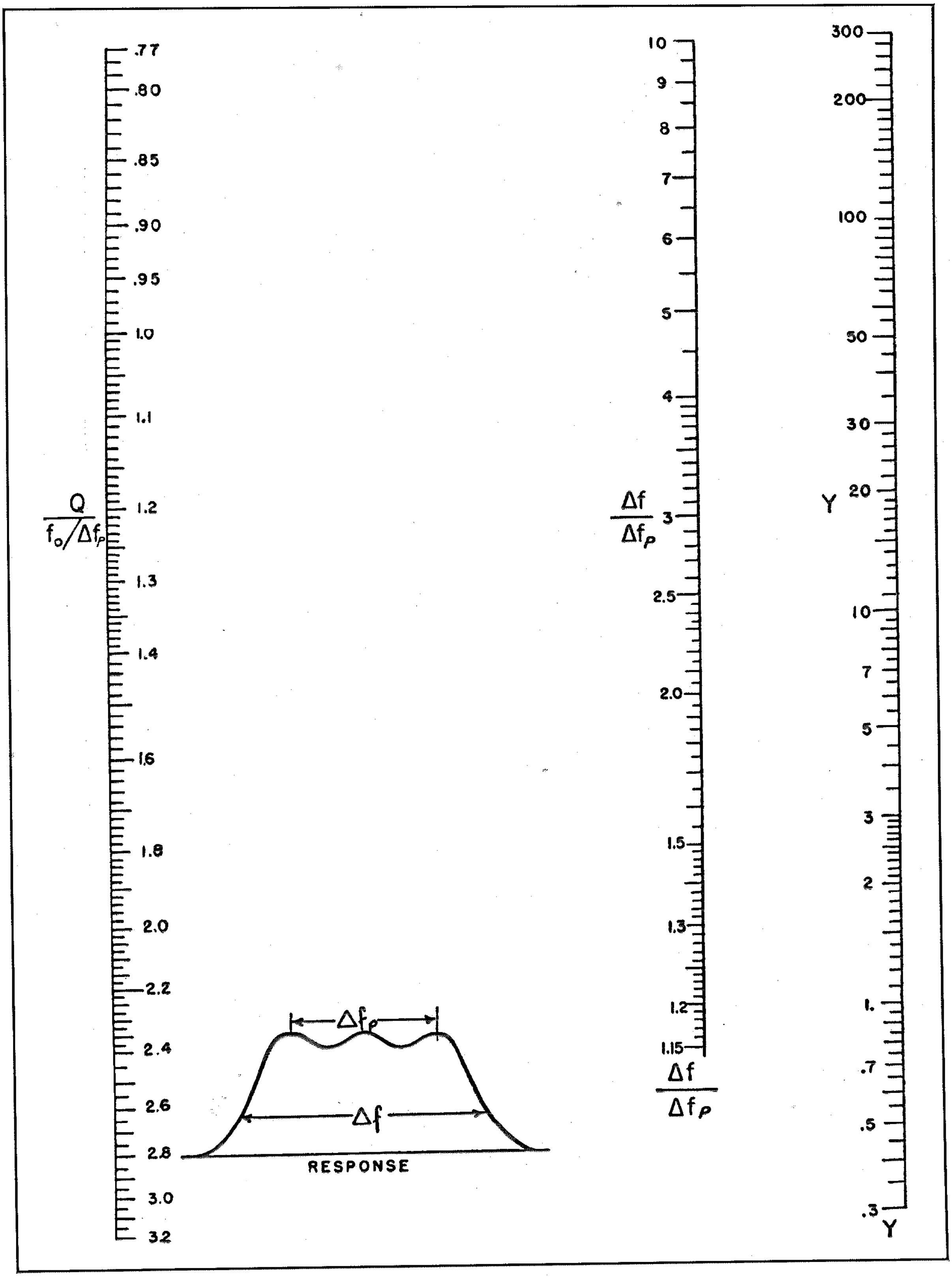

Chart III—Triple-tuned band-pass circuit design.

by the response at the peaks given by (24), we obtain the equation giving the skirt response ratios in terms of the skirt bandwidth.

$$\left(\frac{V_p}{V}\right)_1=\left\{1+\left\{\frac{\left(\frac{\Delta f}{\Delta f_p}\right)\left[\left(\frac{\Delta f}{\Delta f_p}\right)^2-1\right]}{(1+\gamma^2)\gamma}\right\}^2\right\}^{\frac{1}{2}}. \qquad (38)$$

Solution of (38) for $\Delta f/\Delta f_p$ gives

$$\left.\begin{array}{l}\left(\dfrac{\Delta f}{\Delta f_p}\right)=\dfrac{[d+(d^2-4/27)^{\frac{1}{2}}]^{\frac{1}{3}}+[d-(d^2-4/27)^{\frac{1}{2}}]^{\frac{1}{3}}}{\sqrt{2}}\\ \text{where}\\ \qquad d=\gamma(1+\gamma^2)[(V_p/V_v)_1^2-1]^{\frac{1}{2}}.\end{array}\right\} \qquad (39)$$

From this equation we can calculate the skirt bandwidth for different skirt-response points.

The next desired equation is the gain equation. Substituting the condition given by equation (34) into the peak equation (31), we obtain

$$V_2=\frac{1}{\gamma}\times\frac{I}{4\pi\Delta f_p[(C_1+C_{m_1})(C_2+C_{m_2})]^{\frac{1}{2}}} \qquad (40)$$

and, finally, if we substitute the design conditions (given by equations (34) and (36)) in the phase-shift equation (29a), we obtain the phase shift in terms of the peak-to-valley ratio, represented by γ, and the ratio of the bandwidth to the peak bandwidth.

$$\tan\theta=\frac{+\left(\frac{\pm\Delta f}{\Delta f_p}\right)\left[2\gamma^2+1-\left(\frac{\Delta f}{\Delta f_p}\right)^2\right]}{-\gamma\left[\gamma^2+1-2\left(\frac{\Delta f}{\Delta f_p}\right)^2\right]}. \qquad (41)$$

The above equations apply to a single triple-tuned stage. When N stages are cascaded and a resultant peak-to-valley ratio of V_p/V_v is desired, then the peak-to-valley ratio of each stage $(V_p/V_v)_1$ must equal $(V_p/V_v)^{1/N}$. Similar reasoning applies to the skirt-response ratio, so that $(V_p/V)_1=(V_p/V)^{1/N}$.

Application of the above reasoning gives the following design equations for N cascaded triple-tuned circuits, where the input and output resonant circuits in each stage are of equal Q, and the Q of the middle resonant circuit is much higher than that of the input and output circuits.

Let

$$\gamma=\frac{\left\{\frac{1+(V_p/V_v)^{1/N}}{[(V_p/V_v)^{2/N}-1]^{\frac{1}{2}}}\right\}^{\frac{1}{3}}+\left\{\frac{1-(V_p/V_v)^{1/N}}{[(V_p/V_v)^{2/N}-1]^{\frac{1}{2}}}\right\}^{\frac{1}{3}}}{3^{\frac{1}{2}}}; \qquad (35a)$$

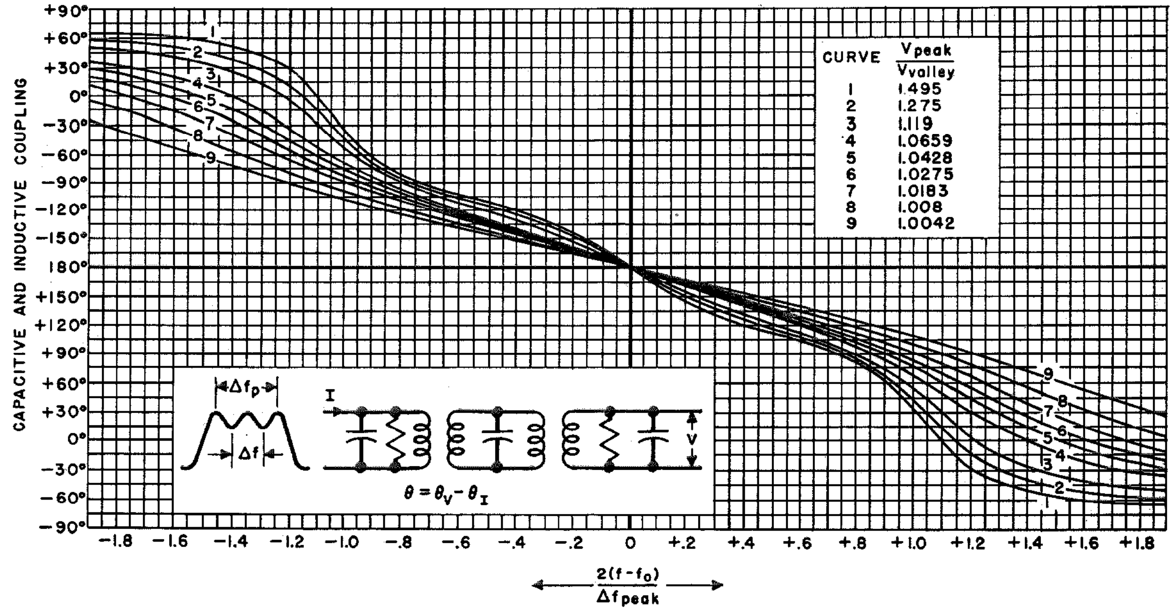

Chart IV—Phase shift for a flat-top triple-tuned circuit ($Q_1=Q_3$, $Q_2\gg Q_{1,3}$) for different peak-to-valley ratios.

then

$$\frac{Q}{f_0/\Delta f_p}=\frac{1}{\gamma}$$

$$KQ=\sqrt{1+1/\gamma^2}$$

$$\frac{V_p}{V}=\left[1+\left\{\frac{\left(\frac{\Delta f}{\Delta f_p}\right)\left[\left(\frac{\Delta f}{\Delta f_p}\right)^2-1\right]}{(1+\gamma^2)\gamma}\right\}^2\right]^{N/2} \tag{38a}$$

$$\left.\begin{aligned}&\frac{\Delta f}{\Delta f_p}=\frac{[d+(d^2-4/27)^{\frac{1}{2}}]^{\frac{1}{3}}+[d-(d^2-4/27)^{\frac{1}{2}}]^{\frac{1}{3}}}{2^{\frac{1}{3}}}\\ &\text{where}\\ &\quad d=\gamma(1+\gamma^2)[(V_p/V)^{2/N}-1]^{\frac{1}{2}}\end{aligned}\right\} \tag{39a}$$

$$\frac{\text{Gain per stage}}{g_m/4\pi\Delta f_p[(C_1+C_{m_1})(C_2+C_{m_2})]^{\frac{1}{2}}}=\frac{1}{\gamma} \tag{40a}$$

$$\tan\theta_{\text{per stage}}=\frac{+\left(\pm\frac{\Delta f}{\Delta f_p}\right)\left[2\gamma^2+1-\left(\frac{\Delta f}{\Delta f_p}\right)^2\right]}{-\gamma\left[\gamma^2+1-2\left(\frac{\Delta f}{\Delta f_p}\right)^2\right]}. \tag{41a}$$

From (34), (35a), (36), (39a), and (40a), another set of nomographs has been prepared. From the phase-shift equation (41), a family of curves has been prepared.

The procedure for using these nomographs and curves is identical with the procedure given in Section 3 for the double-tuned nomographs. The reader should refer to the examples given in that section.

12. Acknowledgment

I wish to express appreciation to R. R. Batcher for the construction of the nomographs from the equations in this paper.

I am also indebted to Jesse LeGrand of the Federal Telecommunication Laboratories for his helpful suggestions as to the format of the nomographs, and to both Mr. LeGrand and Melvin Klein for the many productive discussions we have had about much of the subject matter contained in the paper.

13. Corrections and Notes

Several corrections have been made in this reprint that were not included in the original publication. They occur in equations (15) and (30) and in references to the figures that introduce equations (1), (2), and (3). Also, the fourth line from the end of page 362 now refers to (19) rather than (18).

In Fig. 3, in the denominators of the equations for the vertical legs of the π equivalent of the transformer, $\pm$ should be $\mp$. In the equations for the vertical leg of the T equivalent of the transformer, $\doteq$ should be $=$.

It should be noted that in Fig. 6, Chart A, and equation (27), subscripts 1, 2, and 3 refer to the input, output, and middle circuits, respectively; whereas in equations (28), (29), and (30), they refer to the input, middle, and output circuits, respectively.

Simultaneous Radio Range and Radiotelephone Equipment

By GEORGE T. ROYDEN *

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RADIO RANGE SIGNALS provide reliable indications of the location of an aircraft with respect to a specified course. The range station and the sector in which the aircraft is flying can be identified. Furthermore, voice transmission may be made without interfering with the radio range signals.

The development of radio range operation is reviewed, its theory is discussed, and apparatus is described. This type of equipment serves all important airports not only within the United States but also the principal airways throughout the world.

• • •

The equisignal radio range system[1] employing two crossed loop antennas has been particularly effective for the navigation of aircraft in that only an ordinary receiving set is needed on the airplane. Signals are alternately transmitted from each antenna. The figure-of-eight radiation pattern of the loop antennas produces signals of equal intensity along the bisector of the angle between the two antennas. At any other position, the signal from one of the antennas will be stronger. Usually a dash-dot (*N*) signal will be transmitted on one antenna and a dot-dash (*A*) signal on the other to facilitate identification of the sector when outside of the equal-signal zone. The signals are interlocked and a long dash is heard when on course.

Although the early equipments provided only four fixed courses at 90 degrees with respect to each other, devices were included in later designs for rotating the courses and also for squeezing and bending the courses so they might be useful in the most desirable directions. Subsequently, modulators and change-over relays permitted choice of range or telephone transmissions. This type of equipment is still being employed as localizers for airports of secondary importance and at intermediate points along the airways.

The loop type of radio range was subject to variations, particularly at night.[2] These vagaries, together with the necessity to interrupt the range signals to broadcast weather information and operating instructions by voice, made it desirable to use vertical radiators in lieu of loop antennas and to develop equipment for simultaneous transmission.[3]

After a service trial[4] had demonstrated improved performance, plans were formulated to install the new type of equipment to serve the principal airports in the United States. It is the purpose of this paper to describe, not only the apparatus, but also its function in the system.

1. *Theory of Operation*

The manner in which the simultaneous radio range and radiotelephone system functions is adequately described in text books[5] on the subject and will be reviewed but briefly here.

The radio range station is located several miles from the airfield so its antenna towers will not constitute an accident hazard. Wherever possible, it is aligned so that one course passes over the center of the principal runway. The courses are also aligned so that they are directed along the routes connecting that airport with other airports.

There are five 135-foot vertical tower antennas at each station, four situated at the corners of a square about 425 feet on each side and the fifth at the center. The center antenna is employed for the carrier signal which is modulated by voice. The four corner antennas produce two figure-of-eight radiation patterns.

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* Formerly, Federal Telephone and Radio Corporation, Newark, New Jersey.

[1] F. H. Engel and F. W. Dunmore, "A Directive Type of Radio Beacon and Its Application to Navigation," *Bureau of Standards Scientific Paper 480;* January, 1924.

[2] Haraden Pratt, "Apparent Night Variations With Crossed-Coil Radio Beacons," *Proceedings of the I.R.E.*, v. 16, pp. 652–657; May, 1928.

[3] "Simultaneous Transmission of Voice and Aerial Radio Range Signals," *Air Commerce Bulletin*, v. 5, p. 268; May 15, 1934.

[4] "Simultaneous Transmission of Radio Beacon Signals and Voice in Trial Service at Pittsburgh," *Air Commerce Bulletin*, v. 7, p. 1; July 15, 1935.

[5] P. C. Sandretto, "Aeronautical Radio Engineering," McGraw-Hill Book Company, New York; 1942.

In Fig. 1, there is shown at *C* the circular radiation pattern of the carrier channel produced by the center tower. One figure-of-eight radiation pattern is keyed with a dot-dash sequence as shown at *A*. The other, keyed with a dash-dot sequence, is shown at *N*.

Voice modulation of the carrier channel is accomplished in the normal manner. However, the audio-frequency signal is supplied to the modulator through a preamplifier having a peak-limiting feature and the subsequent audio-frequency amplification is adjusted so that the maximum modulation is 70 percent. The power radiated by the side-band channel on a frequency 1020 cycles per second above that of the carrier channel is adjusted so that 30-percent modulation occurs in a receiver located on the course having the strongest signal.

This division of the available 100 percent modulation is proportioned so that attention may be concentrated either on the voice or the range signals. However, most aircraft are provided with a small filter to separate the voice and range signals so either may be heard separately without interference from the other.

Referring again to Fig. 1, the signals heard in several directions are indicated at the bottom. In the direction where there is negligible radiation of *A* signals, only *N* signals are heard. Next, an off-course signal on the *N* side is represented. Although the tone is continuous, the stronger *N* signals are clearly identified. In the tone where the *A* signals are equal in intensity to the *N* signals, a continuous tone is heard. This zone is approximately 1.5 degrees wide. It is customary for pilots to fly along the right edge of this equisignal zone where the *A* or *N* signal can just be discerned.

The keying of the interlocked *N* and *A* signals is controlled by a motor-driven device which interrupts the radio range signals after 12 *NA* sequences and transmits the coded station identification letters.

2. *Equipment*

The equipment for a simultaneous radio range and radiotelephone station consists of two transmitters (one for standby), coupling unit, five antenna tuning units, transmitter control unit, receiver, five antennas, buildings, power supply, and miscellaneous materials for installation. The principal units are shown in Fig. 2.

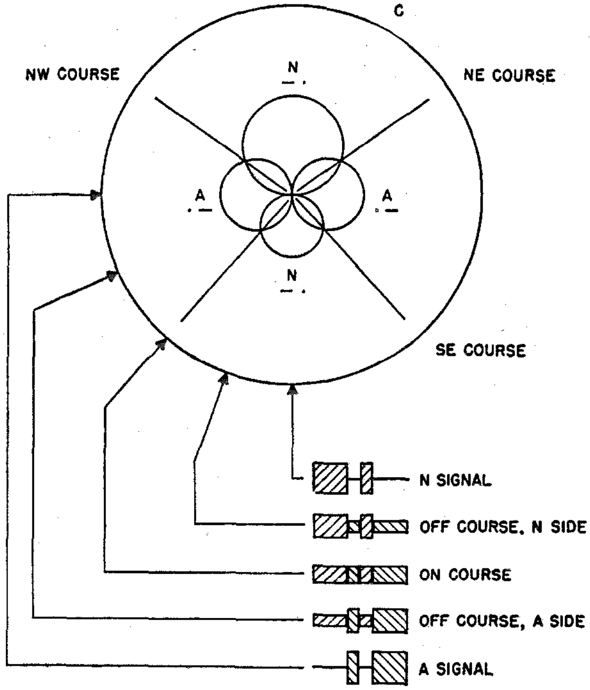

Fig. 1—Typical radiation pattern of a radio range. The asymmetry of radiation in the *N* sectors has shifted the courses from the 90-degree relations that would obtain with equal radiation from each of four antennas at the corners of a square. The center antenna produces the circular field pattern.

3. *Transmitter*

Each of the transmitters, one of which is shown in Fig. 3, has two independent crystal-controlled oscillators, two radio-frequency amplifiers, an audio-frequency amplifier, two rectifiers with filter networks for plate current, one rectifier with filter for grid bias, control switches and relays, indicator lamps, meters, fan, and other necessary auxiliaries.

The oscillator circuits are designed to operate at any frequency between 200 and 400 kilocycles. The oscillator circuit is not tuned to resonance but is designed to provide uniform amplitude throughout the specified frequency range. The crystal for the side-band channel is ground for a frequency 1020 cycles higher than that for the carrier channel. Furthermore, these crystals are matched pairs having similar characteristics so as to maintain this frequency difference over a wide range of temperature.

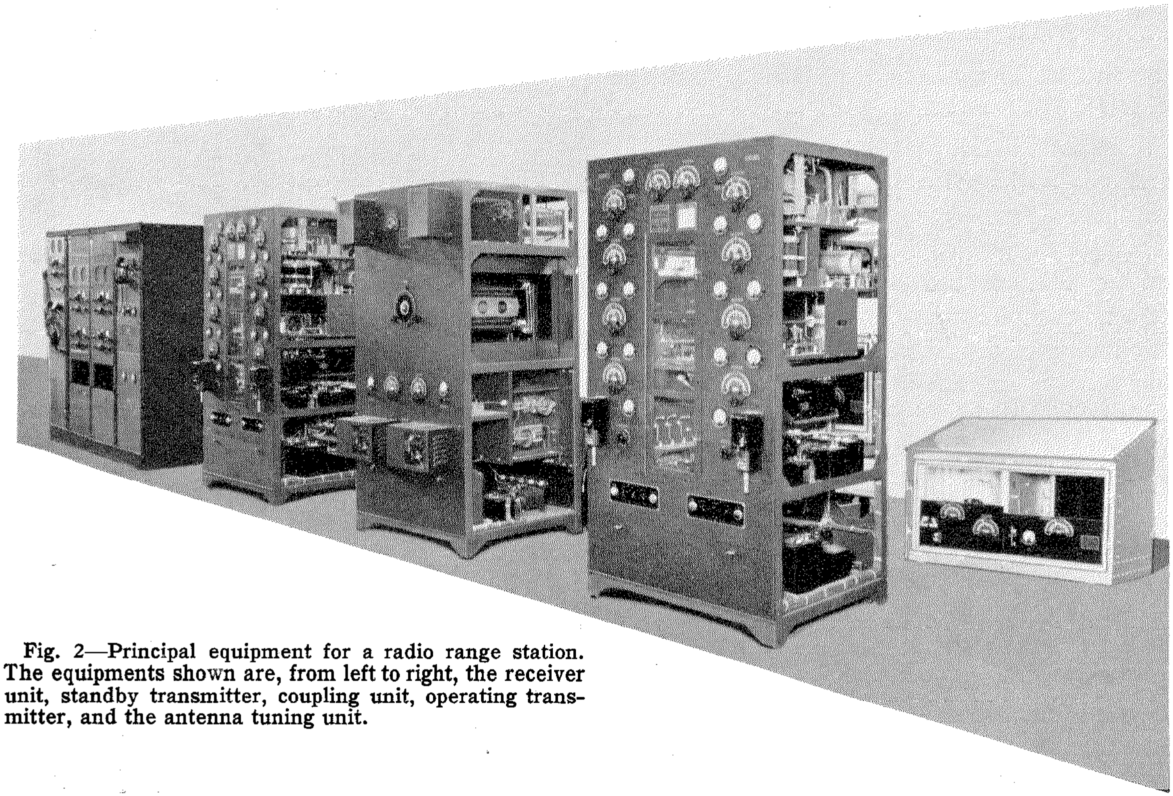

Fig. 2—Principal equipment for a radio range station. The equipments shown are, from left to right, the receiver unit, standby transmitter, coupling unit, operating transmitter, and the antenna tuning unit.

The untuned buffer stage is followed by several tuned stages of radio-frequency amplification. The final stage of the carrier channel employs 4 type-805 triodes in parallel. The radio-frequency amplifier for the side-band channel is similar to that for the carrier channel except that only two tubes are used. The input circuit for the final stage is neutralized. The output circuits are of the π type designed for coupling into a transmission line. Because of the very small frequency separation between the carrier and side-band channels, careful design is necessary to avoid interaction and cross modulation between the two radio-frequency amplifiers.

The audio-frequency amplifier is of conventional design employing three stages with a pair of type-805 tubes operating as a class-B amplifier in the output stage.

One rectifier provides direct current at 1300 volts for the plate circuits of the power tubes. Separate filter networks are provided for the carrier and side-band channels to avoid cross modulation. Another rectifier provides plate current at 500 volts for the oscillator and low-power stages. A third rectifier provides negative grid bias. A thermostatically controlled electric heater ensures proper operating temperature for the mercury-vapor rectifier tubes.

The transmitter is started and stopped by contactors which may be controlled by switches on the front panel or by relays in the transmitter control unit. Appropriate lights indicate the operating conditions.

Controls for the necessary adjustments are mounted on the front panel together with dials. An indicating meter for the controlled circuit is installed in the vicinity of each control.

The inner compartment, in which all vacuum tubes are located, acts as a chimney to provide ventilation. When the ambient temperature is above normal, a thermostat starts a fan to provide additional ventilation.

4. *Coupling Unit*

Specially designed contactors in the coupling unit, shown in Fig. 4, connect either of two transmitters to the transmission lines to the antennas and their tuning units. The side-band energy passes through appropriate keying and power-dividing circuits.

From the relay that selects the proper transmitter, the side-band circuit is connected through the moving contact of the keying relay to the course-squeezing resistors and the goniometer primary windings, which are tuned to resonance with capacitors. The secondary windings of the goniometer are connected through tuning capacitors to the artificial lines and thence to the transmission lines.

The keying relay, which is often called the link-circuit relay because of its position in the circuit coupling the transmitter to the transmission lines, is of special design. The relay is adjusted so that contact is made on one side at the same time contact is broken on the other side. One of each pair of contacts is backed by a spring. This spring is compressed and, on reversal of the signal, assists in accelerating the armature so that a quick break is accomplished.

The course-squeezing resistance pad is connected to a terminal board which has movable links to facilitate connection of the pad in either goniometer primary circuit. The pad may also be connected directly between the link-circuit relay and the goniometer to facilitate adjustment to the required input impedance and to give the proper ratio of currents in the goniometer primaries to produce the desired course squeezing.

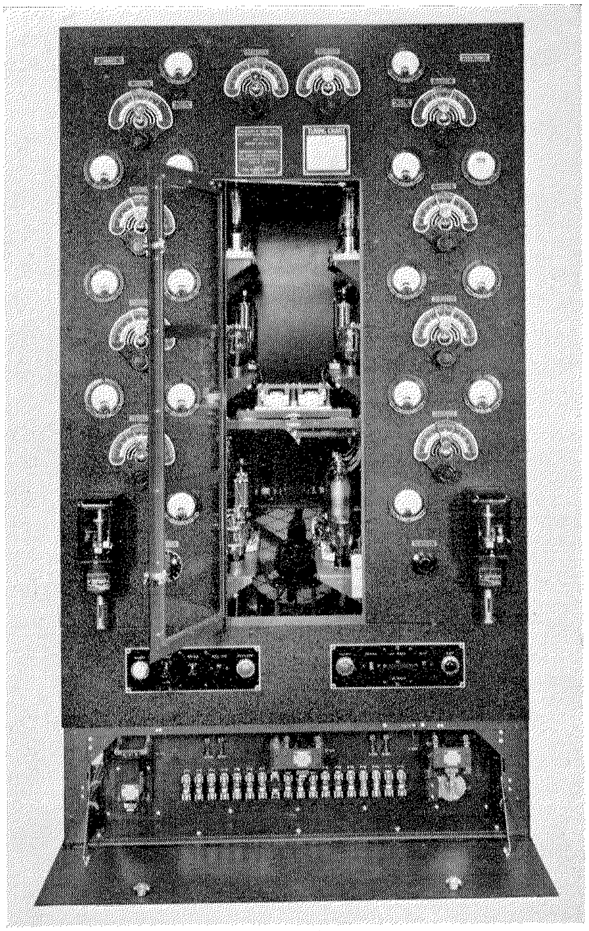

Fig. 3—Radio transmitter. The 400-watt carrier channel is at the left and the 275-watt side-band channel is at the right.

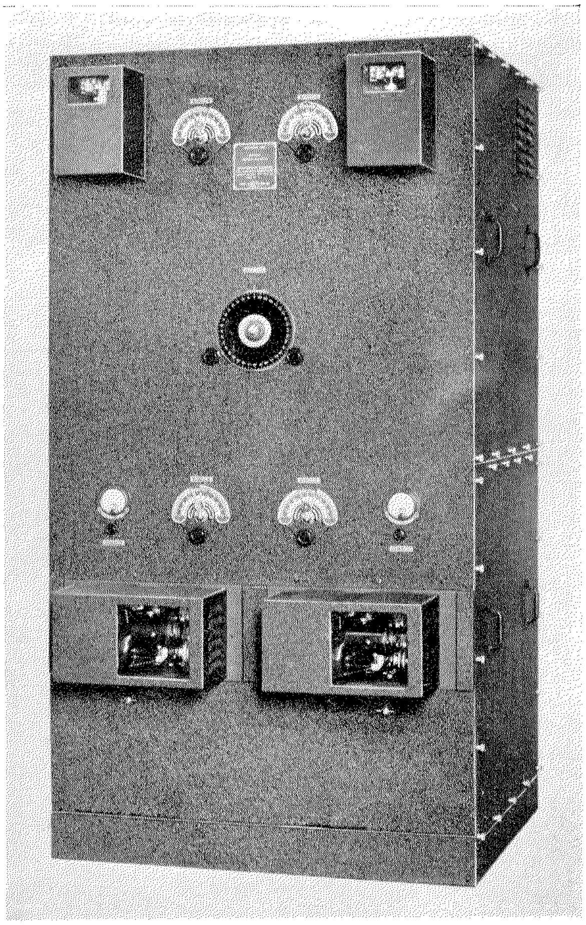

Fig. 4—Coupling unit through which the output power passes from the transmitter to the antenna tuning apparatus. It includes keying relays, goniometer, and other circuits.

The goniometer[6] is an adaptation of the device employed in the early days of radio by Bellini and Tosi in their receiver for direction finding. It was necessary to incorporate many engineering refinements and subtleties of design in the modern version to obtain the high degree of precision required by the exacting specifications. The goniometer has two primary windings at right angles to each other and two secondary windings also at right angles to each other but mounted so that they may be rotated within the primary windings. Each primary consists of two symmetrical windings; the tuning capacitors are inserted between the windings so that the reactive components are electrically balanced with respect to ground. This reduces the effective capacitive coupling, which coupling is further reduced and partially compensated by a shield of special design inside the stator winding. Instead of connecting the wires of this shield in parallel to form a Faraday screen, they are connected in series and so arranged that they carry capacitive currents whose magnetic fields oppose the fields of similar currents in the primary windings. Further compensation is obtained by careful disposition of the leads.

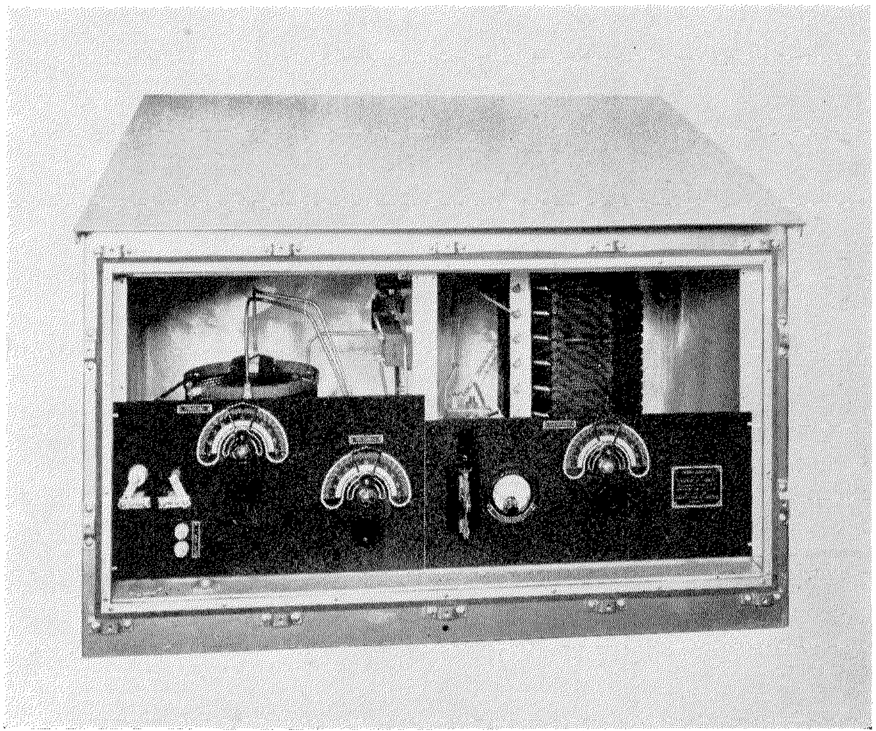

Fig. 5—Antenna tuning unit located at the base of each tower. All necessary antenna adjusting circuits are included in it together with isolation equipment for the tower lights.

A variable air capacitor, shunted by mica capacitors as necessary for the assigned frequency, is employed for tuning each primary and each secondary winding to resonance. Each goniometer secondary circuit supplies energy through two artificial lines and transmission lines to diagonally opposite radiators. These artificial lines are designed so that, in effect, they reduce the length of the real line and are adjustable; the reduction caused by one may exceed that by the other. In this manner, the size of one lobe in the radiation pattern in Fig. 1 may be made larger than that of the other lobe.

Duplicate motor-driven keying devices are provided for controlling the link-circuit relays. Segments may be selected and mounted on a cam to form the coded identification signals. This cam is rotated twice at intervals of about 40 seconds and sends the station call letters once on the *A* lobes and once on the *N* lobes. Keying is accomplished in the side-band transmitter at a low level.

Duplicate rectifiers are included in the coupling unit to provide direct current for operation on the link circuit relays.

5. *Antenna Tuning Unit*

Power is fed from the coupling unit through specially designed flexible underground transmission lines to an antenna tuning unit at the base of each vertical radiator. Each unit has an input circuit, an antenna circuit, and means for coupling them to each other. The input circuit is tuned by a variable air capacitor, shunted by mica capacitors, if necessary. The antenna inductor is tapped and has a variometer for fine tuning.

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[6] W. W. Macalpine, "The Radio Range Goniometer," *Communications*, v. 23, p. 36; December, 1943.

The circuits may be adjusted for phase and amplitude stability[7,8] by making the angle between the antenna current and the exciting voltage equal to the angular length of the transmission line between the source and the connection to the antenna tuning unit. This adjustment may be made by temporarily short-circuiting the transmission line at the source, connecting the line to the primary circuit, tuning the primary circuit (with antenna temporarily disconnected) so that the resultant circuit is in resonance; then adjusting the antenna circuit and coupling so that a nonreactive impedance equal to the surge impedance of the transmission line appears at the input terminals.

Radio-frequency choke coils with by-pass capacitors are included for feeding current to the tower for operation of the obstruction warning lights. The return circuit is through the antenna loading coil to ground. A buried parkway cable brings power from an insulating transformer to the antenna tuning unit at each corner tower.

All electrical components are enclosed in a metal house for protection. A view of the antenna tuning unit is shown in Fig. 5.

6. *Control Unit*

The transmitting equipment is normally unattended and is remotely controlled from an operating position at the airport. The control apparatus is mounted in a cabinet as shown in Fig. 6.

A group of secondary relays, mounted at the top, provide for turning the filaments of either transmitter on or off, energizing the plate circuits of either transmitter, energizing the modulator of the transmitter in use, transmitting an attention signal (series of dashes of 1020-cycle modulation) to warn pilots listening to the range of a forthcoming important voice broadcast, start or stop the marker beacon transmitter, and turn on or off the tower obstruction warning lights.

These secondary relays are controlled by primary relays, which are mounted behind the remote-control panel, located just below the blank panel at the top. The primary relays

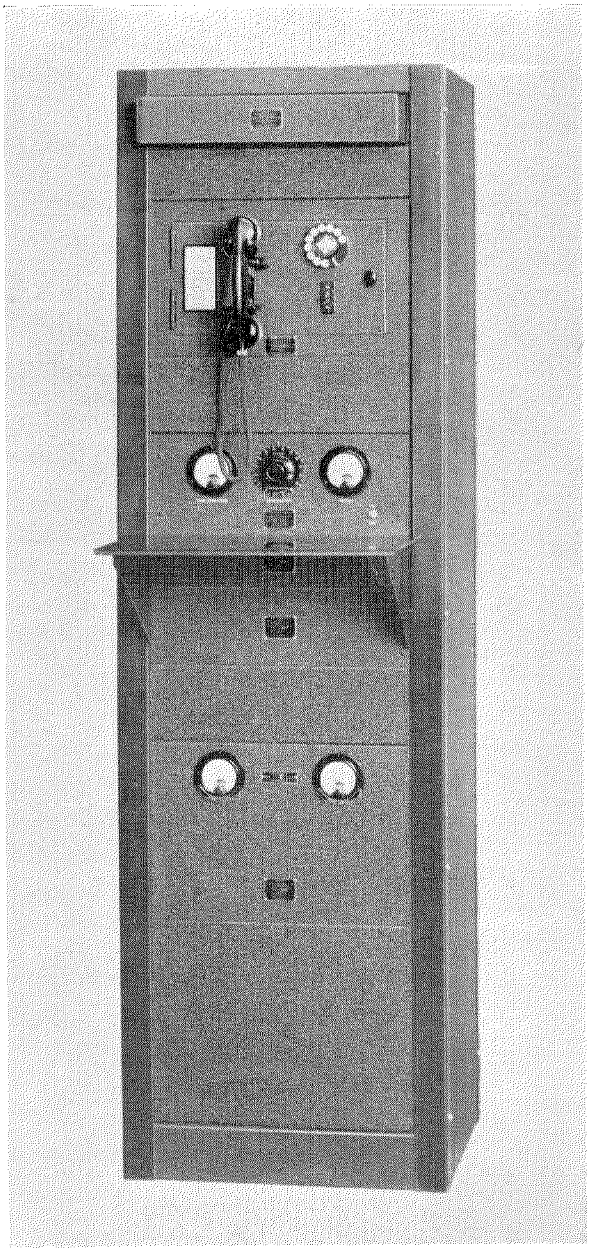

Fig. 6—Control unit permitting remote operation of the unattended transmitting station.

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[7] F. G. Kear, "Phase Synchronization in Directive Antenna Arrays with Particular Application to Radio Range Beacon," *National Bureau of Standards Journal of Research*, v. 11, p. 123; July, 1933.

[8] H. Roder, "Elimination of Phase Shifts Between the Currents in Two Antennas," *Proceedings of the I.R.E.*, v. 22, pp. 374–394; March, 1934.

are controlled by a two-motion machine switch. These are of the type employed with automatic-telephone apparatus. A handset of the sound-power type is provided for communication with the remote station. A dial is mounted on the panel for local control through the remote-control apparatus.

The preamplifier for the voice circuit is mounted near the center of the cabinet. This has sufficient gain to compensate for loss in the transmission line. An important feature is the automatic peak-limiting circuit. A signal proportional to the output is applied to a full-wave rectifier whose cathodes have a positive bias such that only an excessive peak signal will pass through the rectifier and charge a capacitor. The change in voltage, thus produced, biases the grids of two variable-gain amplifier tubes in the first stage of the amplifier.

A band-elimination filter is mounted near the center of the cabinet. This is designed to remove a narrow band of frequencies centered at 1020 cycles from the voice signals to avoid interference with radio range operation.

An adjustable line equalizer is provided to compensate for greater losses at frequencies near 200 and 4000 cycles.

The panel near the bottom provides direct current at 48 volts for operation of the remote-control switch and associated relays. Selenium rectifiers[9,10] are employed in connection with an ingenious circuit to maintain substantially constant output voltage with normal variations of line voltage and load current.

7. *Receiver Unit*

The communication-type receivers are mounted in three adjacent cabinets, as illustrated in Fig. 7. Each receiver may be tuned to a frequency in the 200–400-kilocycle band or to a frequency in the high-frequency band. A loudspeaker is provided for each receiver and is connected through a switch which permits it or all other loudspeakers to be silenced.

A high-quality microphone, mounted on an arm that permits a wide range of positioning, is provided for voice broadcast and radio communication.

A speech amplifier is provided for use in connection with the microphone. It has sufficient gain to raise the level to that necessary for operation over a normal telephone line.

The remote-control panel includes a dial of the automatic-telephone type, for control of the

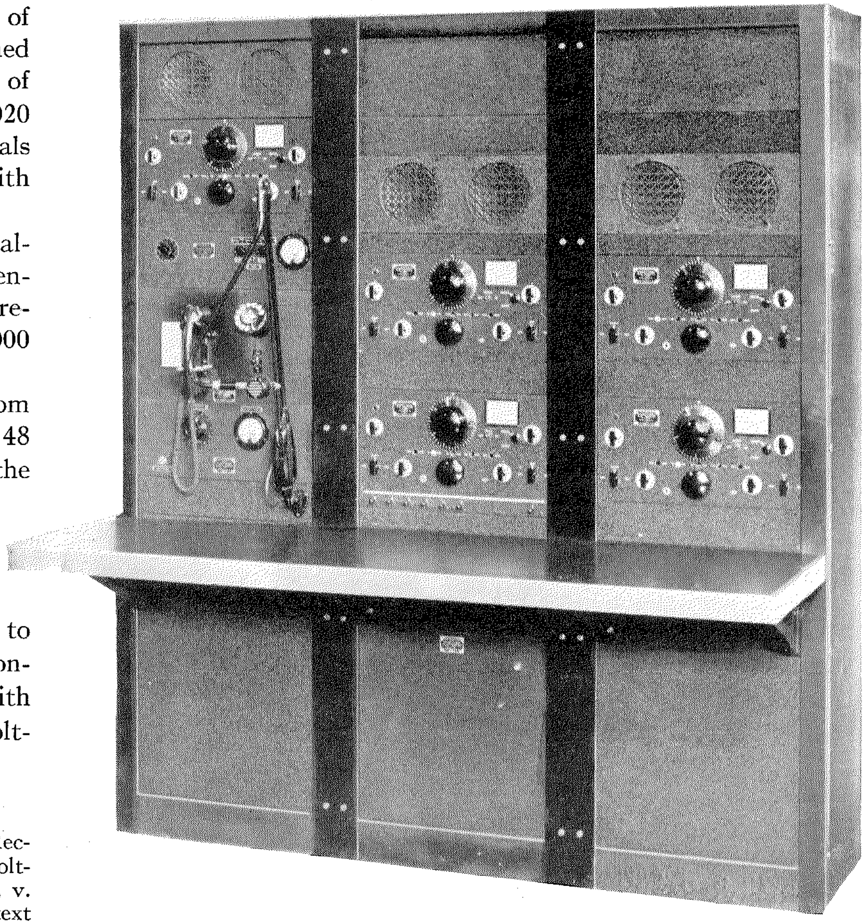

Fig. 7—Several communication-type receivers are mounted in three cabinets. The loudspeakers are arranged so that all but one may be silenced as required.

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[9] J. E. Yarmack, "Selenium Rectifiers for Closely Regulated Voltages," *Electrical Communication*, v. 20, n. 2; Figs 7, 8, and 9, also text on pages 128 and 129; 1941.

[10] U. S. Patent 2,182,666.

apparatus at the transmitting station, and a sound-powered telephone handset for intercommunication.

A monitor panel with a combination band-pass and band-reject filter, meter, and switches is included for observation of the signals from any radio range station in the vicinity.

A shelf is provided for the convenience of the operator.

8. Antennas

The antennas most often used are fabricated steel towers approximately 135 feet tall. They are insulated at the base and operate against a counterpoise about 10 feet above the ground. The use of a counterpoise helps maintain constant capacitance and resistance of the antenna circuit.

9. Production

Although similar apparatus has been made by four other organizations within the United States, the number of equipments manufactured by Federal Telephone and Radio Corporation and its predecessor, Federal Telegraph Company, exceeds the combined production of all other concerns.

10. Acknowledgment

The technical staff of the United States Civil Aeronautics Administration and predecessor organizations were responsible for the over-all system planning. Credit is due to Mr. H. Romander for the design of the transmitter, to Mr. W. W. Macalpine for the design of the coupling unit, to Mr. G. C. Perkins for mechanical design, and to other associates in the Federal Telephone and Radio Corporation who made significant contributions to the design and production of the equipment described.

Variable-Frequency Two-Phase Sine-Wave Generator

By T. H. CLARK and V. F. CLIFFORD

Federal Telecommunication Laboratories, Inc., Nutley, New Jersey

VARIABLE-FREQUENCY two-phase sine-wave voltages have an important application to circle generation in cathode-ray-tube circuits. A generator capable of delivering any frequency between direct current and 60 cycles is described. The sine-wave output may in turn have superimposed on it any frequency from direct current to 400 kilocycles.

* * *

A system of direction finding,[1] which was of great utility during World War II, indicated direction by means of a pattern on the screen of a cathode-ray tube. A simple circle, just within the scale markings of Fig. 1, was obtained in the absence of signals. The radial sweep amplitude was controlled by the output of a radio receiver whose input was derived from a goniometer rotating at the speed of generation of the circle. The field explored by the search coil changes the circle to sharp pointers, as shown in Fig. 1, to indicate the direction of arrival of the radio wave.

1. Rotating Magnetic Field

One method of generating the circle is to rotate an electromagnet about the neck of the oscilloscope. The electromagnet is energized by the plate current of a direct-current amplifier operated from the second detector of the receiver. This is entirely satisfactory for shore or shipboard equipments. However, it is difficult to adapt for aircraft or mobile use because of its bulk and weight.

2. Rotating Electric Field

A circle can also be generated on the screen of a cathode-ray tube by applying a two-phase sine-wave voltage to the deflection plates. Any suitable generator may be used. A two-phase generator or a single-phase generator with a phase-splitter for deriving the orthogonal phase may be employed. The disadvantage of the single-phase generator and phase-splitter is that the phase shift depends somewhat on frequency, so that the two phases may not always be at 90 degrees. Disadvantages of the two-phase generator are that the two phases may not be orthogonal, of equal amplitudes, or perfect sine waves. Filters to eliminate higher harmonics are usually required.

3. Resistive Generator

A resistive sine-wave generator was successfully produced in 1939 in Les Laboratories, Le Matériel Téléphonique in Paris. This generator is shown in Fig. 2. It uses 360 separate resistors in groups of 4, arranged so that their resistance values give a sinusoidal resultant when measured around the circumference of the distributing or commutating cylinder. The commutator consists of copper bars affixed to the ends of a cylinder rotated on its axis and carrying the resistors. Such an arrangement gives perfectly satisfactory operation and long life. It operates at all speeds up to 3600 revolutions per minute and the circle produced on the cathode-ray tube is of excellent

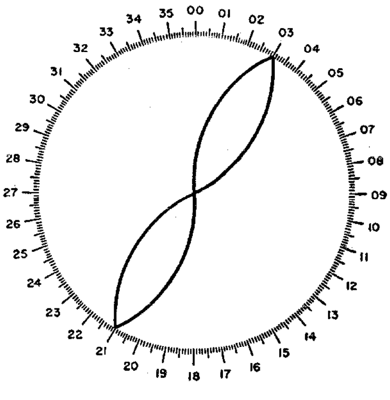

Fig. 1—Indicating pointers on screen of cathode-ray tube show the direction of arrival of incoming waves. In the absence of a signal, a circular pattern is visible just within the scale markings.

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[1] Giltner Twist, "Army Radio D-F Networks," *Electronics*, v. 17, pp. 118–124; November, 1944.

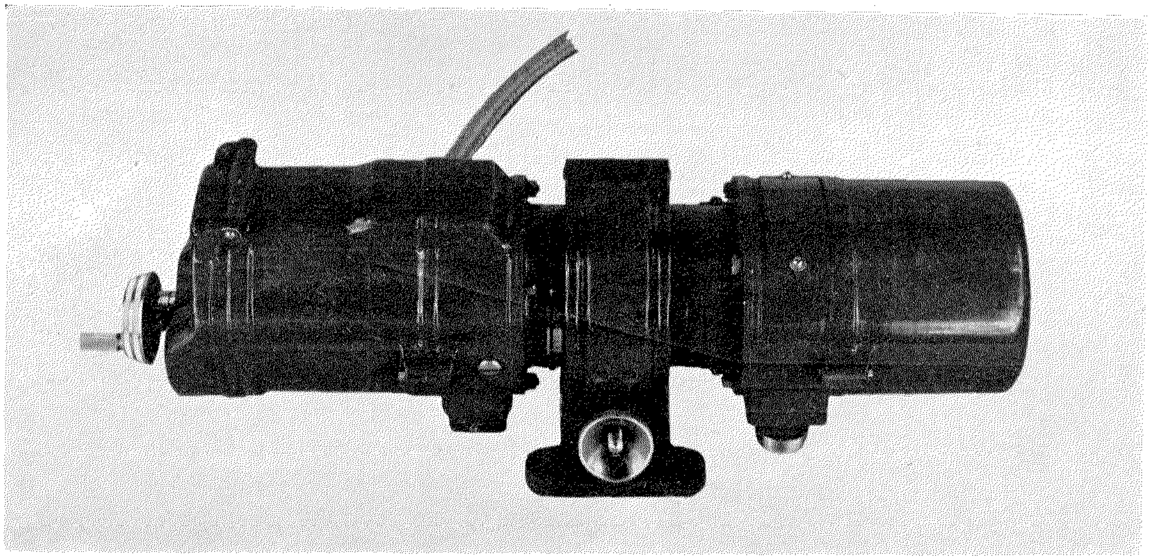

Fig. 2—Resistive type of sine-wave generator produced in 1939 by Le Matériel Téléphonique.

quality. The generator is not useful at high output frequencies unless the resistors (which are wire wound) are noninductive. This adds to the cost.

4. *Linear Distributor*

The unit described in this paper is based on the parallel equipotential lines formed by a uniform flow of current in a plane resistive sheet as shown in Fig. 3. If two probes are used to explore such a resistive sheet, zero voltage will always be found between the probes when they are located on the same equipotential line. If the sheet is rotated about the central point and if the probes are located at the extreme ends of a line passing through that point, the voltage between the probes will vary in a sinusoidal manner. If a second pair of probes is arranged 90 degrees from the first, as shown dotted, a two-phase voltage can be obtained.

The current, which flows through the resistive sheet to produce the equipotential lines, may be derived from any suitable source. The resultant output voltage is easily calculable from the resistance of the strip between the probes and the current flowing through the strip. A schematic diagram of the utilization circuit is shown in Fig. 4, where R is the resistance of the strip, r is the resistance of the load between the probe points, and R_i is the internal resistance of the generator G. The voltage variation across the load resistor

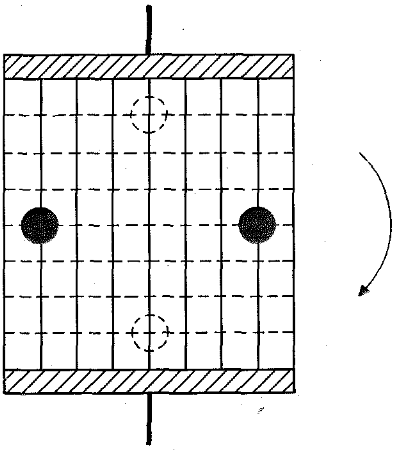

Fig. 3—Linear distributor. Uniform lines of current, indicated by solid lines, flow across a plane resistive sheet from copper connecting bars. Equipotential lines, shown dotted, are at right angles to the current flow. Probes are shown as large solid dots. Relative rotation of the probes about the central point on the sheet will develop a sine-wave voltage between the probes. For two-phase output, the additional pair of probes shown dotted may be used.

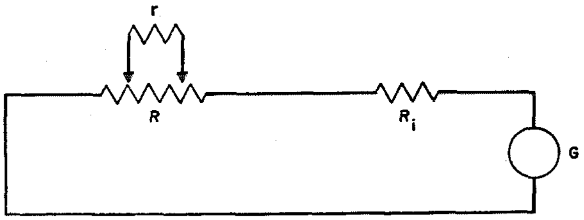

Fig. 4—Equivalent network of distributor circuit. R is the resistance of the strip, r is the load resistance, and R_i is the internal resistance of the generator G.

r will be sinusoidal when *r* is large compared to the strip resistance *R*. This requirement is easily satisfied when the voltage is applied to the deflecting plates of a cathode-ray oscilloscope, i.e., when *r* is the deflecting-plate resistance. A simplified schematic of the circuits used with a direction finder is shown in Fig. 5.

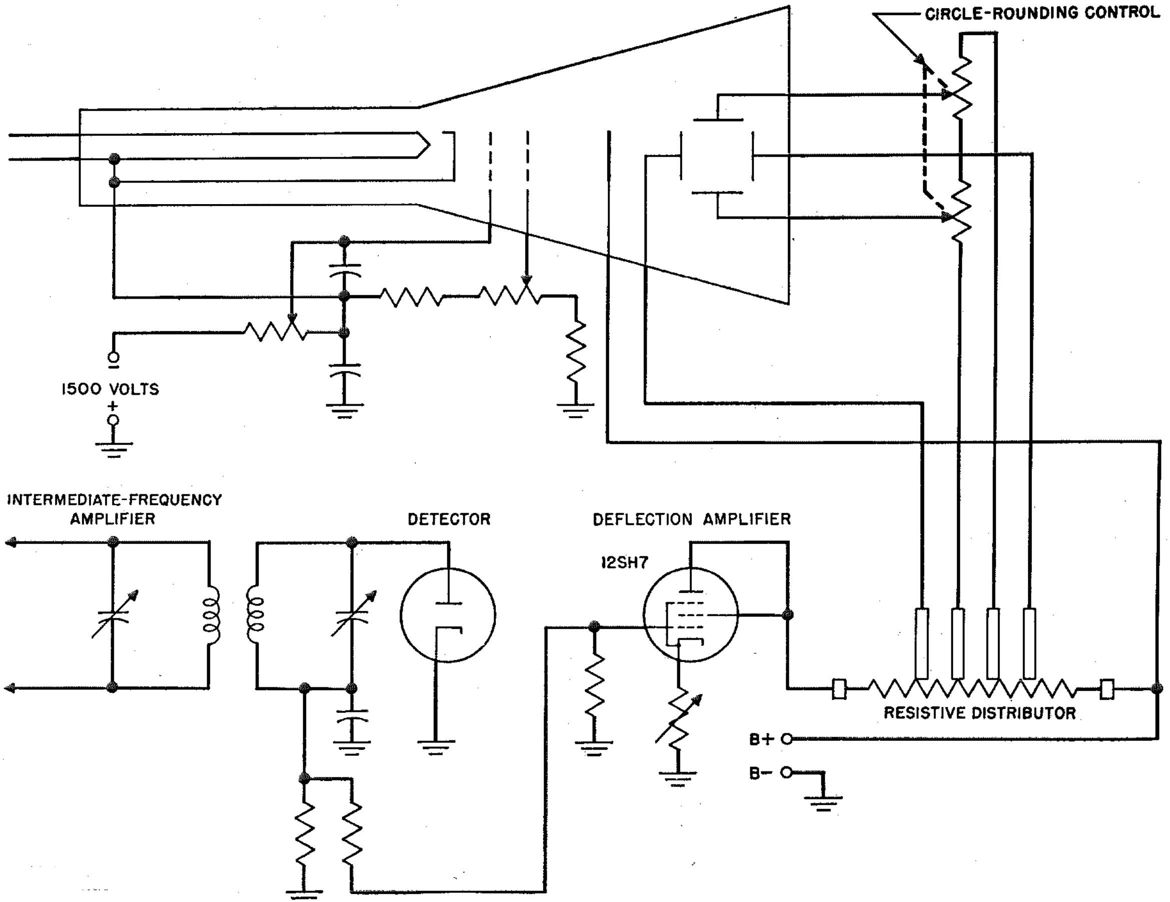

Fig. 5—Application of resistive distributor to a direction finder.

The source need not be direct current. It is satisfactory to pass the plate current of a vacuum-tube amplifier through the resistor strip, in which case the same design considerations should be followed as for any other type of load resistance. The load in the amplifier plate circuit consists of the resistor strip *R* and the load resistance *r*. The associated capacitances and inductances must be considered at high frequencies.

An idea of the circuit design considerations may be obtained from the following application examples. An azimuthal indication accurate to 1 degree is to be produced. The maximum speed of rotation is 1800 revolutions per minute or 10.8×10^3 degrees per second. To obtain the definition required, it is then necessary that the circuit combination shall reproduce frequencies up to 11,000 cycles per second.

In another application, it is desired that 0.5-microsecond square waves be visible on the cathode-ray-tube screen as a radial inward deflection from the circle. For this application, circuit compensation means were employed similar to those in common use in television practice. In the final design, using only the most simple compensation, the deflection amplitude at a frequency of 400 kilocycles was 0.5 of maximum, which was obtained at low audio frequencies. The response was quite flat to the highest frequency. If peaking coils are used, the response can be made better than 0.7 of maximum at 400 kilocycles.

4.1 Distributor as a Constant-Voltage Two-Phase Source

If the rotating resistor strip is energized from a constant direct-current source and the output from the brushes goes to the deflecting plates of a cathode-ray tube, a circle of constant diameter will appear on the screen. If, instead, the outputs of the two phases are applied to direct-current amplifiers, a higher frequency can be superposed on the very-low-frequency two-phase voltages. This method is useful when the two-phase voltage required is greater than can be accommodated safely across the distributor.

The highest frequency that can be superposed on a basic wave depends entirely on the characteristics of the direct-current amplifiers and superposition circuits. Circles have been generated at 150 revolutions per minute and deflected inwardly to zero with 0.5-microsecond square-wave pulses of 5 volts amplitude. The deflecting plates required 1000 volts, peak to peak, at 2.5 cycles with two phases to maintain the circle. The circuit[2] is shown in Fig. 6. It is, of course, very important in applications of this type to maintain a strict balance between the two amplifiers, which must be effective both at direct current and at 3 megacycles.

The pulses are applied simultaneously to the grids of the tubes that perform the superposition. There is one push-pull amplifier for each pair of deflecting plates. 6SK7 tubes are employed for superposition because their remote-cutoff characteristic allows square-law modulation to be used. The output of each superposition stage consists of one phase of the 2.5-cycle voltage on which is superposed the pulse output of the receiver.

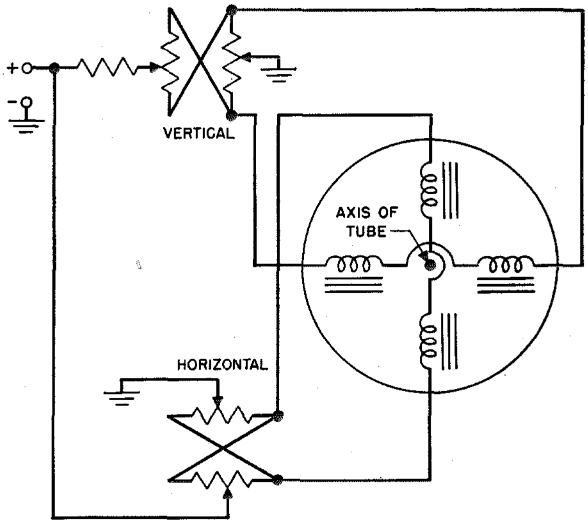

Fig. 7—Horizontal and vertical positioning controls for centering the circular sweep on the cathode-ray-tube screen.

The peak-to-peak voltage of this superposed wave is approximately 150 volts with respect to ground. With this high voltage applied directly to the 6SN7 grids, it is necessary to use a type 9003 pentode as cathode bias resistor. This has the effect of allowing the cathode voltage to follow exactly the large parallel voltages resulting from the superposition process.

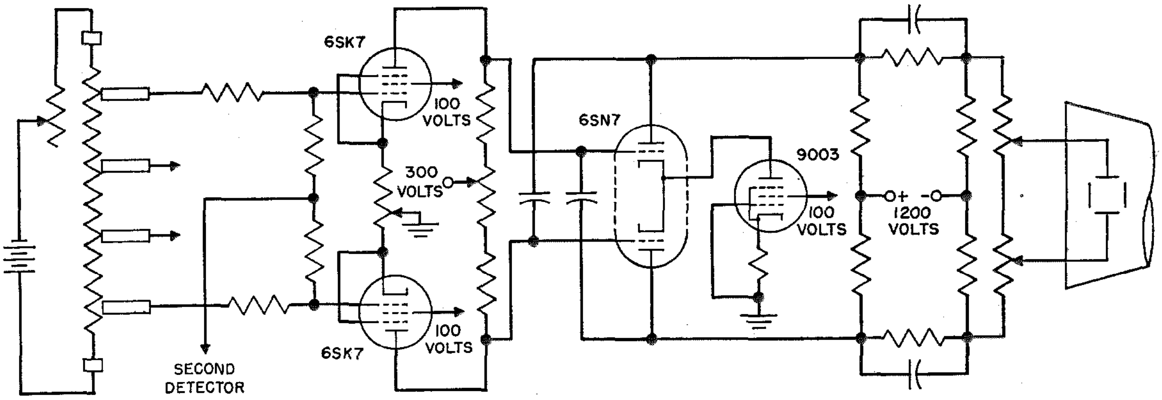

Fig. 6—The voltage from the distributor is amplified and applied to the deflecting plates of the cathode-ray tube. A similar amplifier operates on the quadrature phase from the other distributor probes, supplying voltage to the second pair of deflecting plates and producing a circular sweep. A 5-volt square-wave signal applied to both amplifiers from the second detector of a receiver will reduce the radial deflection to zero. The 9003 is used as a cathode bias resistor for the 6SN7.

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[2] Developed by M. Dishal of Federal Telecommunication Laboratories.

4.2 Means for Centering the Image

The circle, which is generated on the screen of the cathode-ray tube, will be centered only if the gun structure of the tube is centered and if there are no interfering magnetic fields, including that of the earth. These are long-term conditions, and it has been found satisfactory to compensate for them by the application of a magnetic field of controllable magnitude and direction. A suitable circuit is shown in Fig. 7. Adjustment may be made either by centering the position of the circular trace with regard to a calibration scale or by adjusting the spot to the center of the screen or scale when the radial deflection is zero.

A two-phase generator has been designed having only four brushes instead of the six that are required when two balanced phases are desired. The circuit shown in Fig. 8 has been found satisfactory for centering with cathode-ray tubes in which two deflecting plates are connected to the second anode within the tube. The circle produced can be centered along one diagonal of deflection by the voltage divider shown. The supplementary electromagnetic method described above is used to center it along the other diagonal.

5. Development of Distributor

When the idea of a linear resistance strip was proposed, various means were considered for producing a device having long life and reliable characteristics. One proposal was to use a rectangular block of carbon or other resistive material and to electroplate onto it a metal that could be used as a contact and commutating surface. Copper was chosen as the plating metal. The entire device was then machined to produce regular discontinuities in the metallic surface to act as commutating segments. This scheme was unsuccessful because the metallic plating would adhere only by virtue of its continuity. When this continuity was interrupted the metal peeled away from the carbon block.

If an extremely hard resistive surface were available, it was thought that the surface could be used directly for contact with the brushes. A suitable material was produced by the International Resistance Company in the form of a thin carbon film on a bakelite strip. Conductive bars could be plated or otherwise affixed to the terminal ends of the strip. This device was tried with various brushes. It was found that carbon brushes would deposit material on the face of the strip, producing a circle of low resistance and distorting the resultant sweep on the cathode-ray tube. Small metallic brushes consisting of 0.002-inch-diameter tungsten wires were next tried. The resistive material proved so abrasive that the life of such brushes was a matter of hours.

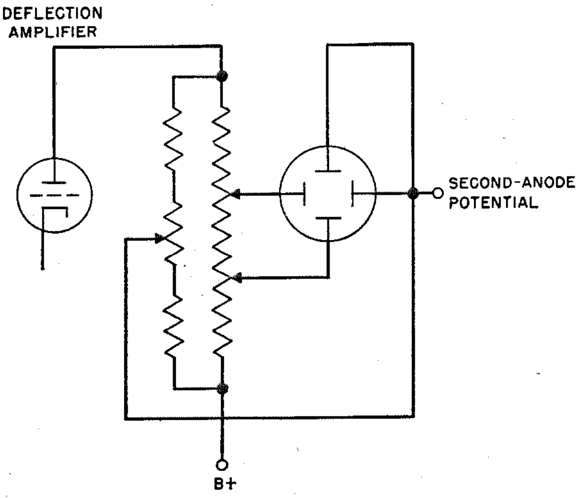

Fig. 8—Centering method used with simplified two-phase generator and tubes having two deflecting plates connected together internally.

Using either of the above resistive materials, it would be possible to wind a conducting wire around the entire block with the windings parallel to the desired equipotential planes. By cutting through each turn on the back of the strip, the resistance between the terminal ends is determined by the resistance of the material between the conducting wires, which serve only as commutating segments. Such a distributor was found to be satisfactory, provided the winding could be kept tight. Various clamping means were used, but fabrication was difficult and this design was abondoned.

Another form of the above design consists of winding a resistive wire on an insulating block. It is then unnecessary to cut the conductors because the resistance is determined by the length of the wire. The resistance between any two points on the same turn will be very small compared with the resistance between the ends of the wire. Using 0.001-inch-diameter Nichrome wire, it was possible to wind a resistor card having adequate resistance for use as a vacuum-tube load. However, bare wire shorts very easily between turns. If insulated wire is used, it is necessary to remove the insulation over the path of the brushes. At the point where the wires are tangential to the circular path of the brushes, the wires are bare for a considerable distance and tend to short between turns. In addition, the brush material deposits between turns of the wire, reducing the resistance, and thereby causing distortion of the circular sweep.

To avoid these difficulties, the wires were covered after winding with a plastic insulating material, which would polymerize on baking. This insulating material was removed to permit brush contact with the conductors. Winding the resistance wire in one plane so that the insulation could be removed without disturbing the position or reducing the diameter of the wire posed a problem that was too difficult of solution and this method was abandoned.

The Formica Company supplied a laminated material made up of alternate layers of phenolic compound and thin copper sheet. A block of such material could be cut to form a commutating device with resistive material applied to one surface and the brushes in contact with the opposite side. Two difficulties presented themselves. First, brushes suitable for use with the copper segments were sufficiently abrasive to pick up the phenolic material, which adhered to them and eventually insulated the brush from the commutating segments. Second, the bond between the phenolic material and the copper was not sufficiently good to allow the production of a mechanically rigid block without additional supports. The material tested was in experimental production only; if perfected, it may be useful for these distributing devices.

6. Final Model

A built-up commutator was designed which eventually proved to be successful. Fig. 9 shows the final model and the commutator assembly is shown in Fig. 10.

The commutator segments are punched from both conducting and insulating strips which are interleaved. The end bars and mounting rods are insulated from the commutator, insulating tubing being used over the mounting rods.

The whole assembly of laminations and two end bars is compressed between locking nuts on the ends of the mounting rods. This assembly is then installed in a jig and ground in a surface grinder so that the face to be in contact with the resistive material is plane and smooth. The resistive material is applied to a card which is placed in intimate contact with the conductive segments. The commutating assembly is then mounted on a base plate so as to compress a resilient backing pad and the resistive material between the base plate and the commutator. This assembly is then tested for contact and linearity of resistance with distance along the commutator.

The entire rotating assembly with the commutator in place is then installed in a jig and the commutator face is ground plane and smooth in the surface grinder. This is necessary because the pressure exerted by the resilient backing pad causes the commutator assembly to bow slightly. If this surface is not plane, the brushes will bounce in their holders. If the surface is not smooth, the brushes will wear and deposit brush material on the insulating segments of the commutator.

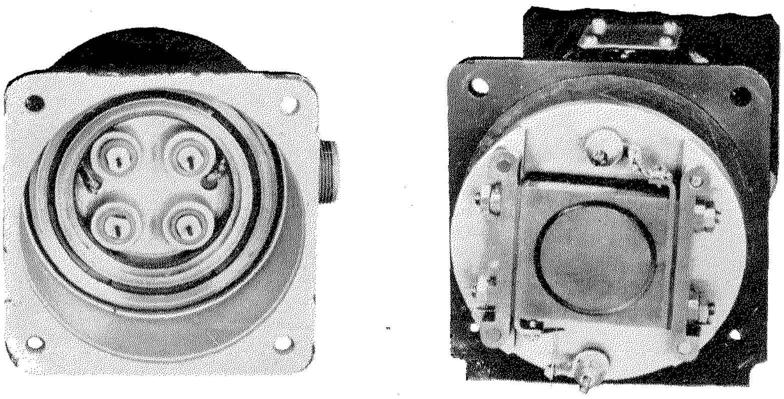

Fig. 9—Final design of commutator-type distributor. The two brushes on the commutator unit make contact with the slip rings on the brush-carrying assembly and provide a path for the input operating current.

In most cases, the base plate is installed directly on the motor shaft. Motors have been procured having end bells machined for accurately positioning the brush holders with respect to the commutating surface. The rotating plate carries two brushes for supplying the current through the resistance strip. These brushes ride on concentric slip rings. This arrangement makes all the brushes contact their slip rings or commutators in the same plane and considerably simplifies the assembly procedure. The mounting means is indicated in the figures. The tandem assembly of a resistive distributor, motor, and goniometer is shown in Fig. 11.

6.1 Design Factors

6.1.1 Vibration

It is necessary that the vibration of the entire assembly and its component parts be kept very small. This requires the surface of the commutator to be at right angles to the axis of the shaft within very close limits. In addition, each rotating assembly is dynamically balanced as a final constructional procedure.

6.1.2 Materials

The slip ring, brush, and commutator materials are chosen for low frictional resistance, low electrical resistance, long life, and reliability during life. This latter point has been one of the most difficult to achieve. A material, which gave good initial operation, often was found to be defective after a relatively short period of operation.

The problems of slip rings, brushes, and commutators have been so omnipresent during the entire history of the electrical age that one would think it should have been completely solved. However, one brush manufacturer stated that over 1800 different types of brushes were manufactured by his organization. The choice of the proper material seemed largely a "cut-and-try" proposition. Consequently, the problem received serious attention.

Commutator segments were made from the following materials: pure copper, phosphor bronze, beryllium copper, stainless steel, Monel metal, Nichrome, pure silver, coin silver, brass, bronze, and aluminum. Brushes in combination with each of the above materials have been tested as follows: graphite (20 different types), copper graph-

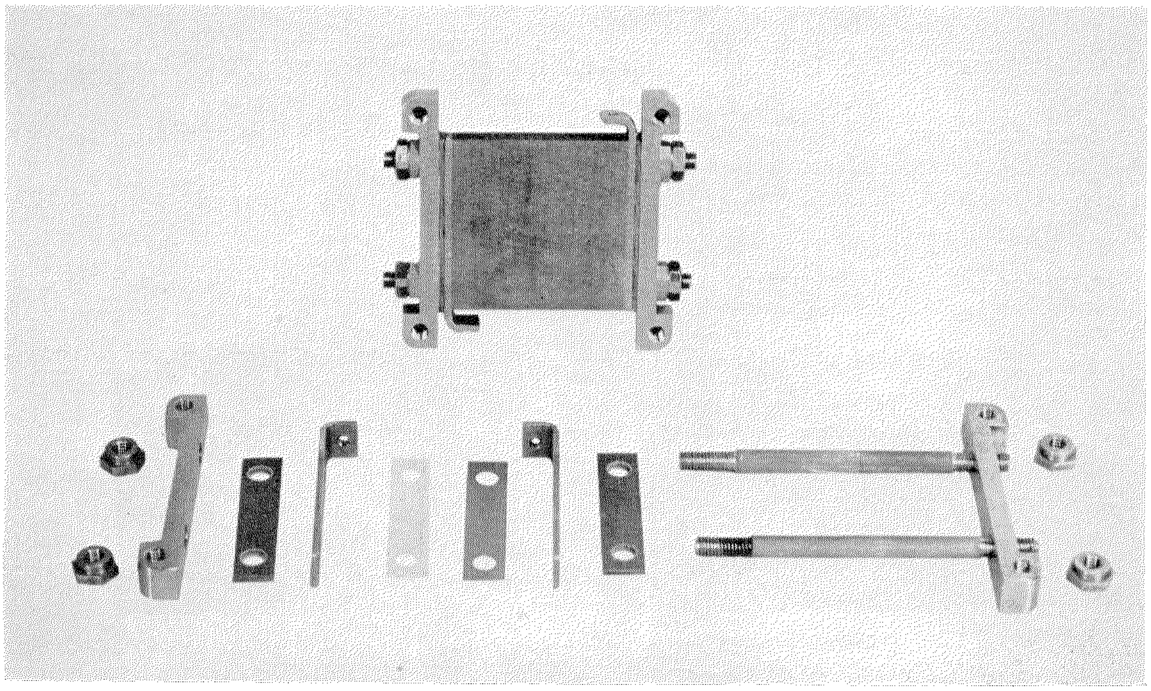

Fig. 10—Commutator assembly. Conducting and insulating laminations alternate. A resistive material makes contact with the conducting sheets on one side while the brushes bear on the other side.

ite, Paliny Number 7, silver graphite, Graphalloy, copper, brass, and Oilite. Insulating materials included: mica, polystyrene, Bakelite, acetate tape, and paper. Many of the above combinations gave devices that would operate more than 1000 hours at 1800 revolutions per minute. The quality of operation was not altogether independent of life and many combinations having long life gave trouble until the brushes had "worn in" or they required cleaning at frequent intervals.

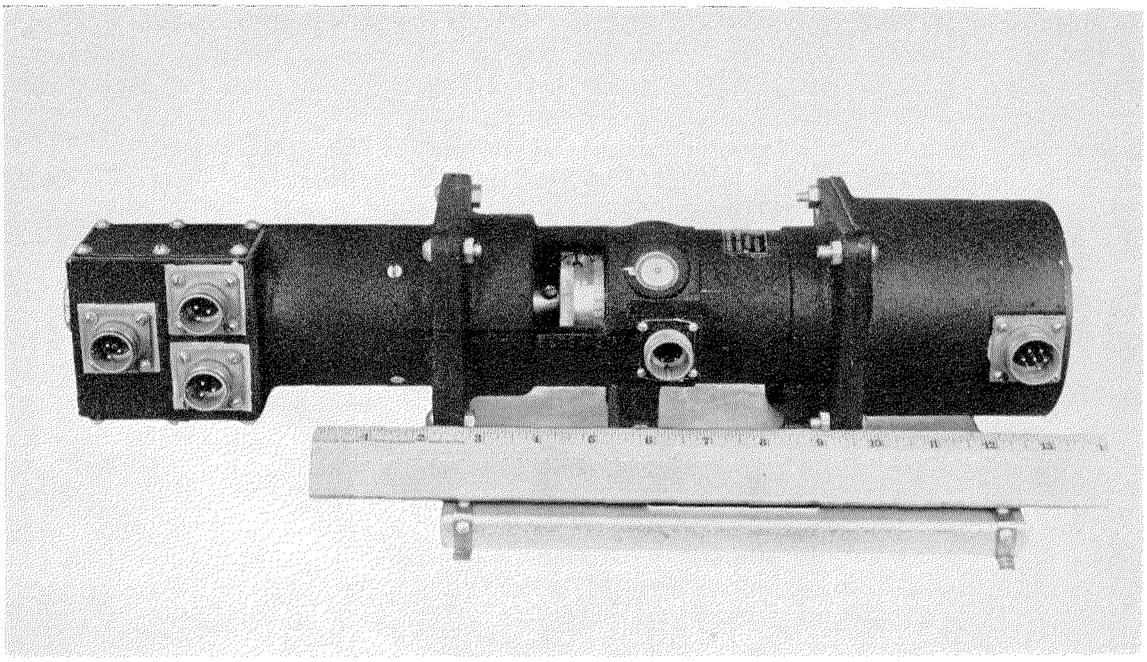

Fig. 11—Assembly of resistive distributor, motor, and goniometer.

The combination eventually chosen consists of a commutator made of alternate layers of coin silver and acetate tape, each 0.005-inch thick. The slip rings are laminated copper and coin silver, silver being the contact surface. The brushes are of grade 411 Silver Graphalloy. With this construction, a number of distributors have operated for more than 1000 hours each without initial difficulties, cleaning, or any maintenance whatsoever. The life of brushes and commutator would appear to be very long.

6.1.3 Springs

Each brush type requires a specially designed spring. Indeed a brush is so closely allied with a particular spring that it has become natural to speak of the assembly as a "brush."

The spring materials tested included: phosphor bronze, music wire, spring brass, stainless steel, and beryllium copper. Beryllium copper was found to possess the most desirable spring characteristics. Correct pressure was determined in each case experimentally. Sufficient pressure to insure contact and yet light enough to prevent abrasion or carbon deposition was the goal.

When the commutator had been made plane so that the brushes did not bounce, and when proper brush and commutator materials had been found, it was relatively easy to find a beryllium-copper coil spring that gave excellent results. The springs used with the slip-ring brushes presented no particular problem.

7. *Conclusion*

The resistive distributor produced has a variety of uses. It can be used simply as a two-phase alternator of variable frequency. It can be mechanically synchronized with any rotating shaft. This is particularly valuable where extremely low frequencies are desired, where constant amplitude is required at variable low frequencies, or where higher-order components or frequencies must be superposed on a basic wave.

It is possible to increase the number of phases simply by changing the machining of the casting that carries the brush holders or probes.

In practical use, the voltage across the strip is limited to 300 volts and no applications have been envisioned where the dissipation in the strip exceeds 5 watts. The resistance of the strip may be varied simply by changing the value of the resistance card. The lowest resistance that has been used is 1000 ohms and the highest, 500,000 ohms. The resistance of the load is dependent on the resistance of the strip and the purity of the wave form desired. There is always a certain amount of commutator ripple, which may be removed by filtering. In most cases, where the voltages are applied to generate a circle in a cathode-ray tube, the ripple is not objectionable.

The device is particularly useful for the generation of two-phase voltages between zero and 20 cycles where, it seems, no other variable-frequency generators of constant output are available. It is also very useful where a distributor of this type is required to have long life at high rotational speeds.

Passenger Telegram Service From Aircraft

Marking a milestone in the history of U. S. A. civil aviation, the Marine Division of Mackay Radio and Telegraph Company, in conjunction with American Overseas Airlines, has inaugurated a new communication service whereby plane passengers may send messages via radiotelegraph from aircraft in flight to any point in the world as well as to ships at sea.

For the present, handling of messages will be confined to transmission from air-to-ground only, with radio communications essential to the aircraft in flight taking priority. Two-way service with American Overseas Airlines is expected to be available in the near future. All messages are now cleared through the Amagansett, Long Island, station of Mackay, which operates on a 24-hour schedule.

As American Overseas planes are staffed with two radio officers, no additional personnel will be required. The standard marine rate applies to these messages.

Dimensional Analysis Applied to Very-High-Frequency Triodes*

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DIMENSIONAL analysis and the laws of similitude are of great service in the study of vacuum tubes. The principle of the method is briefly set forth. It is shown that the operation of similar tubes depends only on a single parameter $\varphi = fd/V^{\frac{1}{2}}$, in which f is the frequency, d a linear dimension of the tube, and V one voltage of the system. The efficiency, gain, Q factor under load, and other dimensionless parameters related to the operation of the tube, can be expressed as functions of φ alone, φ being proportional to the transit angle of the electrons. It is shown, further, that for tubes having the same types of cathodes, the products f^3d and f^4V must remain constant if unvarying efficiency is desired.

Recent improvements in the theory and construction of triodes have extended their utilization by two frequency octaves. A simplified analysis of the motions of electrons in a triode operating class B shows that the properties of a tube at low frequencies are maintained without any substantial deterioration up to values of φ close to 2.5, with $\varphi = fd/V^{\frac{1}{2}}$, f being the frequency in megacycles per second, d the cathode-anode distance in centimeters, and V the anode voltage. For $\varphi = 8$, the gain falls to unity and oscillation efficiency becomes very low. For $\varphi = 2.5$, the Q under load of the output circuit is approximately 18. These results have been checked by experiments.

• • •

A few aspects of the theory of vacuum tubes used as very-high-frequency power generators will be contrasted with receiving amplifier tubes, which are not discussed. This problem raises many difficulties if a rigorous and complete solution is attempted, as the analysis must include the energy exchanges between circuits and electrons moving in electric fields, which vary rapidly over large amplitudes, and in the presence of space charges that are not negligible.

Dimensional analysis has been of great service in numerous technical fields for the study of phenomena not susceptible to complete analysis; it allows generalization of results of limited experiments and associates known theory with realizable experience. In the case of vacuum tubes, it provides knowledge useful not only to the designer of tubes, but also to those utilizing existing tubes.

1. Application of Dimensional Analysis

1.1 REVIEW OF PRINCIPLES

Satisfactory tubes and circuits exist for frequencies up to approximately 100 megacycles per second. Up to these frequencies, powers up to 100 kilowatts may be generated with efficiencies over 60 percent. A fruitful method of extending the frequency range consists in studying what happens with such tubes and circuits when they are used at higher and higher frequencies.

Such a study, called "dimensional analysis" or "mechanical similitude," has been used frequently and fruitfully in engineering. It has produced important results in hydraulics and aerodynamics, in developing theories of turbines and blowers, and in studies of ships and aircraft.

A brief outline follows of the main principle of the method; the reader may consult several treatises[1–3] for more complete information.

Any kind of apparatus is based on the application of a certain number of physical laws, each of which may be expressed as a mathematical relationship (generally differential) between a certain number of physical quantities.

One of these relationships is

$$f(G_1, G_2, \ldots) = 0. \tag{1}$$

The G's are physical quantities, each differing from the others.

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* Reprinted from *L'Onde Electrique*, v. 26, pp. 175–187; May, 1946.

[1] Villat, "Mécanique des Fluides," Gauthier-Villars, Paris, 1938.

[2] Monteil, "Ventilateurs, Soufflantes et Compresseurs centrifuges," Dunod, Paris, 1937.

[3] Riabouchinsky, "Sur Quelques Théorèmes d'Analyse Dimensionnelle," *Comptes rendus Hebdomadaires des Séances de l'Academie des Sciences*, v. 217, pp. 205–208; August 23 1943.

A physical law being a truth independent of the measurement system adopted to express the quantities involved, the mathematical character of function f is necessarily such that (1) is satisfied regardless of the system of units adopted; this is the principle of homogeneity.

The fundamental laws of mechanics show that, within the scope of this branch of science, we are free to choose arbitrarily units for measuring three quantities, the units for the measurement of any other quantity being then defined as functions of these three by the equations of mechanics.

The study of electricity introduces a fourth arbitrary quantity.

The quantities for which the units are arbitrarily chosen are called fundamental quantities, and any other unit is related to the fundamental units by a "dimensional equation," which is the basic equation defining the new quantity.

For example, if $|L|$, $|M|$, and $|T|$ are the fundamental quantities in mechanics, the dimensional equation for a force is

$$|F| = |M| \times |L| \times |T^{-2}|$$

from

$$F = m\gamma$$

and

$$\gamma = \frac{d^2A}{dt^2}.$$

The choice of the fundamental quantities is entirely free so long as their number is observed. We shall use this property for writing dimensional equations with various fundamental quantities selected so as to simplify later reasonings and discussions. Thus, it is permissible to change the system of units without changing the meaning of an equation such as (1). For example, (1) may be stated as the relationship between factors $P_1, P_2, P_3, \ldots.$

$$\varphi(P_1, P_2, P_3, \ldots) = 0. \tag{2}$$

Each of the factors P is of the form

$$P = G_1^{\psi_1} \times G_2^{\psi_2} \times G_3^{\psi_3} + \ldots.$$

Then if the dimension of P is expressed by means of dimensional equations relating G to the fundamental quantities A, B, C, and D,

$$P = A^{\alpha} \times B^{\beta} \times C^{\gamma} \times D^{\delta},$$

and of necessity

$$\alpha = \beta = \gamma = \delta = 0.$$

In other words, any physical property can be expressed by means of a function φ of parameters P. The P parameters are *dimensionless numbers*, whose numerical values are unaffected by a change in the system of units. This theorem expresses only the coherence of the reasoning of theoretical physics and has, consequently, a great value of synthesis. One more step only has to be taken to apply it to the theory of similar systems.

Consider a physical system for which certain properties are defined by equations relating dimensionless numbers:

$$P_1, P_2, \ldots, P_n.$$

Consider a second system whose corresponding dimensionless numbers are

$$P_1', P_2', \ldots, P_n'.$$

If

$$P_1 = P_1', P_2 = P_2', \ldots, P_n = P_n',$$

the two systems are said to be similar with respect to the properties studied and all the corresponding equations are identical in pairs.

Any number such as $Q = f(P_1, P_2, \ldots, P_n)$ has the same value in two similar systems. If Q is a number that cannot be found theoretically, but can be obtained experimently, the value of physical similarity is that a single experiment giving Q for one system also gives a knowledge of Q for the infinity of *all similar systems*. This method has considerable value either when direct experiment is impracticable as in the study of ships' hulls or in obtaining information of a general nature from a single experiment as in the case of turbines.

The method of similitude is powerful and in the theoretical field leads to very elegant proofs. However, any good geometrician knows that elegance is of value only in expert hands and wrong steps are all the easier as the reasoning is increasingly abridged. It should always be carefully ascertained that the use of similitude is justified for a particular study. It must be remembered that often physical properties exist that will not be included in the scope of systems that are similar in other respects. By way of example, consider two resonant cavities of similar geometry. Let $K = |L_2|/|L_1|$ be the ratio of similitude of the dimensions. What is the ratio of the resonant frequencies?

The electrical resonance of a cavity is a phenomenon determined by wave propagation and, in particular, by wave velocity C. Now the velocity C is constant for all cavities; giving the relationship

$$v=\frac{|L|}{|T|}=\text{constant.}$$

Thus

$$\frac{|T_2|}{|T_1|}=\frac{|L_2|}{|L_1|}=K \quad \text{or} \quad |f|=\frac{1}{|T|}.$$

This gives the well-known relationship between resonant frequency and geometrical dimensions.

$$\frac{f_2}{f_1}=\frac{|L_1|}{|L_2|}=\frac{1}{K}.$$

The voltage magnification factor Q, being a dimensionless number, might be considered to be the same for all cavities. This is not the case, even assuming all the cavities to be of the same metal, copper, for instance. In such a case, the metal constituting the walls of all cavities has the *same resistivity* but this experimental condition is outside the scope of our hypotheses on the similitude of cavities.

A study of this problem shows that Q is proportional to $\lambda^{\frac{1}{2}}$.

Despite the very real difficulties of this kind and the necessity for careful thinking, the method of similitude has shown itself very fruitful in the investigation of ultra-high-frequency power tubes, a theoretical analysis of which has not yet been made and which involve rather delicate experimental technics.

1.2 Fundamental Hypotheses

The essential elements in one of the systems to be discussed are the cathode, empty space for the motion of electrons, and associated resonant cavities.

For a first simplified study, the following hypotheses and assumptions are made concerning these elements.

1.2.1 Cathode

The cathode consists of a metal surface whose chemical nature and temperature are such that the electron flow leaving it is *limited by space charge.* This condition is fulfilled for all current densities A between zero and a fixed value called "maximum cathode emission." This maximum emission A, in amperes per square centimeter, is a characteristic of the type of cathode.

It is essential for the validity of the following calculations to assume that in no other point of the tube there exists an emissive surface obeying another law than space-charge limitation. This implies that no point of the cathode is saturated, and that there is no secondary emission at the points of impact of the primary electrons on an electrode.

1.2.2 Resonant Cavities

The resonant cavities are assumed to be made of copper. They thus have the properties discussed as examples in Section 1.1. The fact that these cavities are made of copper introduces conditions of primary importance relative to resistive losses.

This prevents the inclusion of oscillating cavities in a similitude system used for analysing electron ballistics in the tube. Accordingly, separate consideration must always be given to:

A. Oscillating cavities, as being the seats of electric currents and resistive losses, and

B. Empty space, which shall be called "tube," in which electrons move between electrodes made of a perfectly conducting material.

Only resistive losses appear in the *cavities*, and only losses by electron bombardment appear in the *tube*. This distinguishes between "tubes" and "circuits" for theoretical purposes.

This distinction can always be made, though in practice it constantly happens that a single surface element often belongs both to the tube and to the circuit. Nevertheless, this distinction is useful for understanding phenomena, for effecting calculations, and for separating losses.

Between tubes and oscillating cavities, relationships exist that are included under the title "coupling of electron beam with resonant cavities."

1.2.3 Empty Space in Which Electrons Move (Tube)

The electrons move freely in empty space within the tube under the action of existing fields. Their trajectories are from the cathode to a final electrode, called the anode from which there must be no secondary emission.

For the purposes of calculation, the following assumptions are made:

A. The effect of magnetic fields on the motion of electrons is negligible. This, for the time being, rules out the case of the magnetron. Similarly, the effect of the magnetic fields generated by the motion of the electrons themselves is neglected. These effects are small, numerically, and have been neglected also in the articles of Llewellyn and Brillouin.

B. The dimensions of the tube are sufficiently small with respect to wavelength to neglect the propagation time of the fields. In other words, the calculated potentials generated by the charges are assumed to differ negligibly from the actual retarded potentials.

C. The potential differences involved are large enough so that the effects of the initial velocities of the electrons are negligible. This assumes that the potentials used are, in general, larger than *10 volts.*

D. The potential differences involved are small enough to make unnecessary a correction for relativity. This means that the electron potentials must not exceed 50,000 to 100,000 volts.

1.3 SIMILITUDE EQUATIONS FOR VACUUM TUBES

Two vacuum tubes may be considered to be similar and to oscillate in similar conditions when:

A. Geometries of the two tubes are similar,

B. Laws of variation of the currents and voltages with time are the same, the period being taken as a unit of time, and

C. At corresponding instants of the cycle, the distribution patterns of the potentials inside the tube are similar.

Under such conditions, it is necessary only to write the homogeneous equations governing the tube operation to obtain very valuable information. There are three such equations.
Poisson's equation:

$$\Delta V = 4\pi\rho.$$

Equation for the motion of an electron in an electric field:

$$m\vec{\gamma} = e\vec{\mathcal{E}}.$$

Limiting condition *at the surface of the cathode:*

$$\vec{\mathcal{E}} \leqslant 0.$$

The fact that the electric field cannot be positive at the cathode surface expresses the hypothesis that the current emitted is limited by the space charge. This condition is independent of frequency.

The dimensionless products P of these equations, relative to various systems of fundamental quantities, may then be formed. The theory of similitude states that for similar tubes, operating under similar oscillation conditions, these numbers P and any dimensionless number are the same.

1.3.1 Choice of Fundamental Quantities

We shall choose as fundamental quantities those whose values are maintained constant by the very nature of the problem or whose variations can be fixed by the conditions of our study.

The first class involves the ratio e/m. In all vacuum tubes, only one kind of electrons move, and these have a constant ratio e/m. In all systems of fundamental quantities, we shall always choose $|e/m|$, which is a simple numerical constant, as a first quantity. This reduces to three the number of fundamental quantities whose units can be changed.

Current density $|A|$ has a maximum value imposed by the physiochemical constitution of the cathode, and can thus be introduced in a simple manner.

The quantities whose value we are free to choose are:

$|L|$, defining the dimensions of the tube, we shall designate by d to avoid any confusion with inductance;

$|f| = 1/|T|$, defining the frequency of oscillation; the main parameter in our study;

$|V|$, defining the scale of potentials inside the tube. With e/m and f common, there remain three fundamental systems:

$$\left.\begin{array}{llll} \frac{e}{m} & f & d & V, \\ \frac{e}{m} & f & d & A, \\ \frac{e}{m} & f & A & V. \end{array}\right\} \qquad (3)$$

1.3.2 Poisson's Equation

In a vacuum where the electrons move, this equation is written

$$\Delta V = 4\pi\rho.$$

This equation, which defines the relationship between potentials and charges in the tube, can be written in dimensional notation as

$$Q=\frac{Vf}{Ad}, \qquad (4)$$

Q being the first dimensionless product. The relationship between the charge density per unit volume and the current density per unit area is

$$A=\rho\frac{dx}{dt}$$

or, in dimensional notation,

$$A=\rho\times d\times f.$$

Besides, $\Delta V=V\cdot d^{-2}$; hence, substituting in Poisson's relationship, (4) is obtained.

It is useful to note that product Q is precisely the Q factor under load of the output cavity of the tube. A is proportional to the convection current of the electrons and, therefore, to the active current through this cavity. Vf/d is the current density through a capacitor, having a dielectric thickness d, under the action of a potential difference V at frequency f. But Q under load is the ratio of the reactive to the active current in the oscillating circuit.

Thus, any variation of Q in the operating conditions of a tube is accompanied by a proportional variation of Q factor under load, of the output cavity.

1.3.3 Equation of Motion

$$m\vec{\gamma}=e\vec{\mathcal{E}}.$$

The equation of motion can be written immediately in dimensional notation as

$$\phi=\frac{m}{e}\frac{f^2d^2}{V}. \qquad (5)$$

V or d must be eliminated between (4) and (5) to obtain the dimensionless products corresponding to the systems of fundamental quantities in (3). Rewriting them according to (5), the following three relationships are obtained:

$$\phi=\frac{m}{e}\frac{f^2d^2}{V}, \qquad (5)$$

$$\phi Q=\frac{m}{e}\frac{f^3d}{A}, \qquad (6)$$

$$\phi Q^2=\frac{m}{e}\frac{f^4V}{A^2}. \qquad (7)$$

Finally, it is interesting to eliminate f between (4) and (5) so as to obtain a relationship giving the current density A independently of frequency. This relationship, which is independent of frequency, expresses the limiting conditions at the cathode surface.

$$\left.\begin{aligned}\frac{Q^2}{\phi}&=\frac{e}{m}\frac{V^3}{A^2d^4}\\ \text{or}\qquad \frac{Q}{\phi^{\frac{1}{2}}}&=\left(\frac{e}{m}\right)^{\frac{1}{2}}\frac{V^{\frac{3}{2}}}{Ad^2}.\end{aligned}\right\} \qquad (8)$$

This is the Child-Langmuir relationship in its most general form and we may note that Langmuir[4] established it by means of dimensional analysis as early as 1913.

Relationships (5), (6), (7), and (8) contain considerable information, which we shall now try to express in a less subtle form.

We note, first, that e/m being an absolute constant, there are four parameters for describing the operating conditions of a tube: f, d, V, and A. Among these parameters there are three relationships: Poisson's equation (4), equation of motion (5), and, finally, the condition: $\mathcal{E}=0$ at the cathode surface, expressing the hypothesis that the emission is always limited by space charge.

Consequently, the problem depends on *only one independent variable.* This first piece of information simplifies considerably the study of a family of similar tubes. We have become accustomed in numerical applications to the use of the variable

$$\varphi=\left(\frac{e}{m}\phi\right)^{\frac{1}{2}},$$

$$\varphi=\frac{fd}{V^{\frac{1}{2}}},$$

f being the frequency in megacycles, d a dimension of the tube (the cathode-anode distance in a triode, for instance) expressed in centimeters, and V a voltage, such as the plate voltage, expressed in volts. The importance of this variable was pointed out in 1940 by Mr. Moisson at Les Laboratoires, Le Matériel Téléphonique, in Paris. In practical applications, φ varies between zero and a few units. It is easily determined that φ is proportional to the "transit angle" often used in the literature on this subject, though its exact definition is generally somewhat vague.

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[4] Langmuir, "The Effect of Space Charge and Residual Gas on Thermionic Current in High Vacuum," *Physical Review*, Series 2, v. 11, p. 450; December, 1913.

From the above considerations, the following results have been obtained in a family of tubes, geometrically similar, when the electrical oscillating conditions are similar:

A. The dimensionless products ϕ, Q, and their combinations have the same numerical values for all tubes of the family and for all operating conditions.

B. Any other dimensionless number expressing a property of the circuit within the scope of the hypotheses of similitude also has the same numerical value for all the tubes considered; such is the case, particularly, for the tube efficiency η and for the power-amplification factor G.

In other words, the Q factor under load, η, and G are single-valued functions of parameter φ. The same thing holds for ϕQ and ϕQ^2.

It is only necessary, therefore, experimentally or theoretically, to establish the relationships

$$Q = f_1(\varphi),$$

$$\eta = f_2(\varphi),$$

and

$$G = f_3(\varphi)$$

to be able to predict the values of these functions for any tube of the family at any voltage or frequency.

As already stated this increases considerably the extent of information acquired by a given experiment. Some practical applications of this theorem will be given to confirm the theory completely.

A consideration of the products ϕQ and ϕQ^2 adds some useful information. For a constant value of φ, therefore of the efficiency η, the products

$$\phi Q = \frac{m}{e}\frac{f^3 d}{A} \quad \text{and} \quad \phi Q^2 = \frac{m}{e}\frac{f^4 V}{A^2}$$

are unchanged numerically.

If we assume all the tubes of the family have the same type of cathode, the value of maximum emission A is constant. Thus, to keep the efficiency constant, $f^3 d$ and $f^4 V$ must be kept constant.

This proves two propositions of major importance in the theory of tubes for very high frequencies. If it is desired to build similar tubes for operation at higher and higher frequencies f with a constant efficiency η, and if all the tubes have the same type of cathodes,

A. All the linear dimensions of the tubes must be reduced proportionally to $1/f^3$.

B. All the voltages applied to the tubes must be reduced proportionally to $1/f^4$.

These two theorems have particularly unfortunate consequences and indicate the difficulty of designing tubes for high frequencies. When a tube design approaches its limit of utilization, even a reduction to microscopic dimensions allows only small increases in frequency. Further, the power that can be generated vanishes abruptly in the neighborhood of the limiting frequency of the family because the linear dimensions vary as f^{-3} and the voltages as f^{-4}.

Because the areas of the tube decrease rapidly with linear dimensions, the power decreases between f^{-4} and f^{-10}. A complete expression of the factors ϕQ, ϕQ^2 also shows that a gain may be obtained for operation at extreme frequencies by an increase in cathode emission A. Improvement of cathodes is, consequently, an important means of extending the field of use of any type of tube except magnetrons at high frequencies.

Relationships (6) and (7) play an important part in the study of ultra-high-frequency tubes, and their complete meaning must be well understood to avoid experimental efforts whose uselessness can be foreseen.

1.4 Q FACTOR, CIRCUIT LOSSES

In the above discussion, the tube alone has been considered, in contrast to the circuits. The resistive losses in the copper circuits must not be included in the scope of the laws of similitude used for the tube. Thus the term "tube efficiency" previously used must be understood as being the ratio of the power delivered by the tube to the circuit to the power delivered by the direct-current source. The only losses reducing efficiency are the losses by bombardment resulting from the residual velocity of the electrons at the time of their final impact. Resistive losses in the circuits have not yet been considered.

Obviously, these resistive losses are very important because, in general, we are interested not in the power delivered by the tube to the circuit, but in the power delivered by the circuit to the external load.

Here, again, dimensional analysis and (8) can give some information on the subject. Rewrite

(8) in the form

$$\frac{Q}{\varphi}=\frac{V^{\frac{3}{2}}}{Ad^2}. \tag{9}$$

The above discussion showed that Q/φ, a dimensionless number, is a function of φ alone. It may also be shown that Q/φ is a constant, independent of φ for any value of φ between zero and a certain limit. When proving Langmuir's law for direct currents, i.e., for $f=0$, we find that

$$\frac{Q}{\varphi}=\frac{V^{\frac{3}{2}}}{Ad^2}$$

has a finite value, which is Langmuir's constant.

If the tube is oscillating at a frequency f, sufficiently low so that at any instant in the cycle the relationship between the instantaneous values of A and V are the same as for direct current, Q/φ has the same numerical value as for $f=0$. Such is the case so long as the transit angles are small and there is no appreciable difference between the actual velocity of the electrons, expressed in volts, and the value of their potential. When the frequency increases enough for the potentials to undergo important changes during the transit time of the electrons, the convection currents A have a tendency to decrease below the value calculated for the steady state, and Q/φ increases. The relationship between Q and φ is thus represented by a curve of the shape shown in Fig. 1.

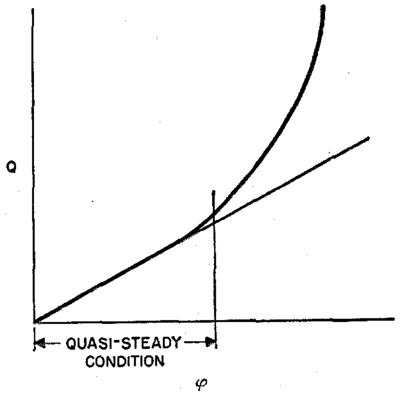

Fig. 1—Q the voltage magnification factor of a loaded circuit is plotted against φ.

The result is that the Q under load of the output cavity of the tube increases more rapidly than φ. Now if Q_v is the Q of the output cavity in the unloaded condition, the efficiency of power transfer from the tube to the load resistance is $1-Q/Q_v$. For good efficiency, Q/Q_v must be kept small; otherwise an important fraction of the power is dissipated in resistive losses in the copper of the cavity.

To operate a tube with an increased φ, which is an easy method for increasing f, it is necessary to associate it with cavities having high Q_v's in the unloaded condition. This stresses the necessity for a very thorough study of the resonant cavities associated with the tube. There is a practical limit to the Q_v's that can be obtained with copper cavities. Moreover, the maximum value of Q_v *decreases* when the frequency increases.

For triodes, tetrodes, and klystrons in which all the energy is collected in a single output cavity, the resistive losses in the circuits are always almost of the same importance as the losses by bombardment inside the tube. It is necessary, therefore, for frequencies higher than 300 megacycles, to give the greatest attention to the practical design of the cavities and to limit the losses strictly to the unavoidable resistances of the copper.

Above a certain frequency, varying between 1000 and 3000 megacycles according to the type of tube, it becomes impossible to obtain good efficiency because of the chemical nature of the metal constituting the output circuit and of that of the emissive surface of the cathode.

There are only two possibilities of extending the frequency limit of vacuum tubes.

A. An increase in current density. This has been done in pulse tubes and explains the successes achieved in this field.

B. Use of multiple-cavity systems which increase considerably the frequencies at which the difficulties appear.

Point B will be considered later as it explains the enormous advantages shown experimentally by magnetrons.

To sum up, dimensional analysis has yielded the following information:

A. The results obtained with a family of similar tubes depend only on one parameter φ.

B. To increase f, while keeping φ at a reasonably low value, d and V must be decreased at a very rapid rate as their effect is proportional to $1/f^n$.

C. The increase of Q under load of the circuits limits the use of tubes with a high value of φ.

Fig. 6 shows curves of efficiency as a function of φ for various triodes used as oscillators. These curves, from an article[5] published in 1945, are very incomplete because of a scarcity of available information.

2. *Electron Transit Time in Triodes*

Neglecting the diode, the triode is the oldest and simplest vacuum tube. It is formed of two spaces, the cathode-grid space and the grid-plate space. The cathode-grid space provides for modulation and the grid-plate space serves both for acceleration and output.

According to recent views, the tube is used by associating each of the two spaces with a resonant cavity which, as far as practicable, is a volume of revolution. Such an arrangement is quite different from the classical conception of a grid circuit and a plate circuit. However, the new conception is to be preferred for an analysis of the tube and circuits for very high frequencies.

Although, in the last few years, changes in construction of active electrodes have been very limited, considerable progress has been made in modifying the conductors connecting the electrodes to the rest of the circuits. Until recent years, the connectors were metal rods of small diameter passing through the glass as in the construction of incandescent lamps. The operation of the tube was based on the idea of "conducting wire"; this is completely obsolete for high frequencies.

The use of circular seals of copper or of a metal having a low temperature coefficient of expansion has made possible the insertion of triodes in revolution cavities. Powers and efficiencies obtained five or six years ago at a given frequency are obtained with this newer technic at frequencies four times greater. This is true for the entire scale of powers.

The smallest triodes now deliver a few deciwatts at 3000 megacycles compared to the same power obtained with acorn tubes at 700 megacycles. Recent tubes can deliver 100 kilowatts at 100 megacycles. Such power was not obtainable before the war above 25 megacycles. This considerable progress indicates the importance of investigating the principles that made it possible.

2.1 Modulation Space

The cathode-grid space is subjected to an alternating field on which an unvarying direct-current field is sometimes superimposed.

When the unvarying field is zero, class-B modulation occurs and this operation is very often used for generating high frequencies. The superposition of a retarding unvarying field leads to class-C modulation. Class A is not used when high efficiencies are desired.

In class-B operation, the field in the modulation space is purely alternating. During the positive half-cycle, emission takes place at the cathode; during the negative half cycle there is no emission and the field at the surface of the cathode is negative.

A general study of the phenomena involved has not yet been published. Llewellyn's calculations, limited to class A and to small amplitudes, do not apply. A first attempt by Wang[6] indicated the principle of a method of diagraming electron motions. A more-recent article by Brillouin[7] yielded very interesting results, but cannot be used at the highest practical frequencies.

2.1.1 *Quasi-Steady State*

The value of current emitted per square centimeter of cathode area is given by Langmuir's equation

$$A = 2.33 \cdot 10^{-6} \frac{V^{\frac{3}{2}}}{d^2} \cdot$$

V is the alternating potential of the grid, d is the cathode-grid distance, and $V = V_E \sin \omega t$.

The desired output modulation is thus produced. But, further, the electrons reaching the grid have a variable velocity corresponding to the variable grid potential. The superimposition of these two modulations gives rise to a large power consumption at the modulation source. This power can easily be calculated

$$dw = A\,V\,dt = 2.33 \cdot 10^{-6} \frac{V_E^{5/2}}{d^2} \sin^{5/2} \omega t\, dt.$$

---

[5] Lehmann and Vallarino, "Study of Ultra-High-Frequency Tubes by Dimensional Analysis," *Proceedings of the I.R.E.*, v. 33, pp. 663–665; October, 1945.

[6] Wang, "Large-Signal High-Frequency Electronics of Thermionic Vacuum Tubes," *Proceedings of the I.R.E.*, v. 29, pp. 200–213; April, 1941.

[7] Brillouin, "Transit Time and Space-Charge in a Plane Diode," *Electrical Communication*, v. 22, n. 2, pp. 110–123; 1944.

Whence

$$W=2.33\cdot 10^{-6}\frac{V_E^{5/2}}{d^2}\cdot\frac{1}{\pi}\int_0^{\pi}\sin^{5/2}\theta\,d\theta,$$

or, finally, after integration, in watts per square centimeter

$$W=10^{-6}\frac{V_E^{5/2}}{d^2}, \text{ approximately.}$$

This is a somewhat academic expression, assuming that the tube has a perfect theoretical cathode. It is very important to note that the excitation power is the power necessary for exciting an inverted amplifier; as the source of excitation, which establishes the potential difference between cathode and grid, has to deliver the total cathode current.

We finally reach the conclusion that the normal mode of excitation of a triode is between cathode and grid, with the considerable concomitant power absorption. This results from the conception of a cavity, by virtue of which the phenomena in the grid-plate cavity are independent of the phenomena in the cathode-grid cavity. These two are completely separated by the grid, except for the cathode-plate capacitance C, the effects of which will be studied elsewhere. Of course, all the energy delivered by the source of excitation to the modulation space exists as kinetic energy in the electrons at their exit from this space. At that moment, the electrons going through the grid enter the output space and the energy of excitation they bring with them will have to be included in the energy balance for this space. Limiting ourselves to the quasi-steady state, two main cases are of immediate interest.

First case: There is no alternating voltage in the output space $V_O=0$. In such a case, no exchange of high-frequency energy can take place in this region, and $W_P=W_B+W_E$, where W_P is the power dissipated by bombardment, W_B is the power delivered by the direct-current source of acceleration, and W_E is the power of excitation. The excitation power is lost completely and exclusively by bombardment of the anode.

Second case: There exists in the output space an alternating voltage such that the plate is brought to a constant direct voltage V_B with respect to the cathode; this condition may be written

$$V_O=-V_E,$$

V_O being the useful oscillating voltage and V_E the oscillating excitation voltage.

In this case, the electrons reach the anode with a constant velocity corresponding to V_B and the power dissipated is only that delivered by the acceleration source.

$$W_P=W_B=I_B\cdot V_B.$$

We thus have, in this case, $W_O=-W_E$.

In the output space, power equal to that *delivered* to the modulation space is collected. At low and medium frequencies, a circuit arrangement called "grid excitation" permits the power developed in the output space to return to the modulation circuit. Consequently, the power dissipated by excitation is zero; this is the result of compensation (often involuntary), through the use of positive feedback, the existence of which is shown by analysis from the ultra-high-frequency viewpoint. In many cases, when the transit times are not infinitely small, this compensation does not take place, a phenomenon difficult to interpret by other theories. The compensation circuit is shown in Fig. 2. A careful examination of these views with a frankly critical mind concerning the conceptions generally used at medium frequencies is recommended.

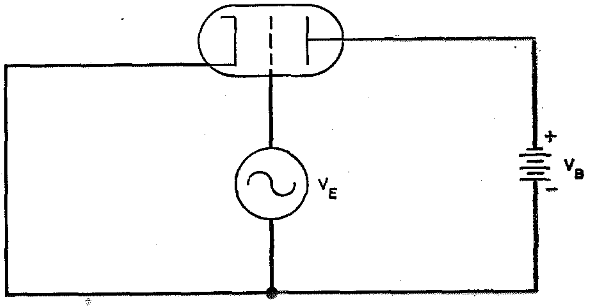

Fig. 2—This circuit arrangement, commonly used at medium and low frequencies, provides an inherent positive feedback that reduces the power required for excitation.

Fig. 2 shows that the grid-cathode circuit is completed through a *wire* that is also part of the output circuit. The impedance of this connection constitutes the positive feedback path that vanishes if a two-cavity circuit is employed as in Fig. 3. The ultra-high-frequency technique *requires* the use of cavities, and wire circuits are completely ignored. Normal excitation between cathode and grid will, therefore, always include the corresponding power consumption. In some circuits, specially designed for high frequencies, a reduction of apparent excitation power is obtained by appropriate feedback.

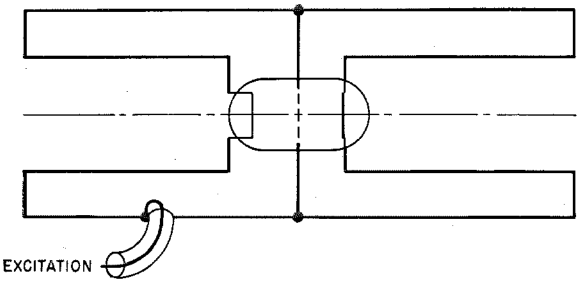

Fig. 3—Circuit employed at high frequencies. Resonant cavities for the cathode-grid and grid-anode spaces eliminate the positive feedback which characterizes the circuit of Fig. 2.

2.1.2 Transit Angles Not Negligible

No complete solution[6,7] of the problem has been given for the case where the transit angle is not negligible.

To obtain in a simple manner a qualitative indication of the phenomena involved, consider class-*B* operation but neglect space charge. This assumption is not justified but it is useful for an understanding of the phenomena. The numerical coefficients obtained, it is true, are fairly different from the exact values.

Let $\theta=2\pi t/T$ be a variable representing time. A given electron is characterized by angle $\theta=\tau$ corresponding to the time when it leaves the cathode with a zero velocity.

θ being the independent variable and x the ordinate as measured from the cathode, τ is the parameter characteristic of an electron.

We shall calculate the law of motion of the electrons assuming a fixed value for the alternating field

$$\mathcal{E}_E=\frac{V_E}{d}.$$

The differential equation for the motion is

$$m\frac{d^2x}{dt^2}=e\mathcal{E}_E\sin\omega t.$$

Integrating twice and writing $v=0$ and $x=0$ for $\tau=0$, we obtain:

$$v=\frac{e}{m}\frac{\mathcal{E}_E}{\omega}(\cos\tau-\cos\theta),$$

$$x=\frac{e}{m}\frac{\mathcal{E}_E}{\omega^2}[\sin\tau-\sin\theta+(\theta-\tau)\cos\tau].$$

This equation can be represented by a single graph, which can be used in all possible cases, by writing

$$\frac{x}{d}=\frac{e}{m}\frac{V_E}{4\pi^2f^2d^2}[\sin\tau-\sin\theta+(\theta-\tau)\cos\tau].$$

Let

$$\varphi_E=\frac{fd}{V_E^{\frac{1}{2}}}.$$

If f is expressed in megacycles, d in centimeters, and V_E in volts, we have

$$\left(\frac{2e}{m}\right)^{\frac{1}{2}}=6\cdot10^7$$

whence

$$\varphi_E^2\frac{x}{d}=45[\sin\tau-\sin\theta+(\theta-\tau)\cos\tau]. \quad (10)$$

Fig. 4 shows the corresponding curves $y=f(\theta,\tau)$ letting

$$y=\varphi_E^2\frac{x}{d}.$$

These curves may be used as follows.

Axis 0θ represents the cathode. After selecting the value of φ_E to be studied, a horizontal line $y=\varphi_E^2$ is drawn. This corresponds to $x/d=1$ and represents the grid.

The portion of the diagram under this horizontal line represents the motion of the electrons between cathode and grid. The following essential points are to be noted.

The electrons emitted between $\tau=0$ and $\tau=\pi/2$ never fall back on the cathode. They always reach the grid. If $\varphi_E<9.5$, they reach the grid during the cycle in which they have been emitted; if $\varphi_E>9.5$, they may oscillate in the cathode-grid space during several cycles. Obviously, this phenomenon is very troublesome and it seems that

$\varphi_E=9.5$ constitutes the maximum practical value.

The electrons emitted between $\pi/2$ and π reach the grid during the first cycle or fall back on the cathode, increasing its temperature by bombardment. It will be noticed that if $\varphi_E<9.5$, the angle during which the electrons go through the grid is smaller than π. The effect of transit time is, thus:

A. To change operating conditions from class B to class C.

B. To cause a fraction of the emitted electrons to fall back on the cathode; this fraction may be as much as 50 percent.

The more rigorous study of Brillouin leads to the same conclusions.

Thus, the effect of transit time in the modulation space is:

A. It cannot, in itself, explain a lowering of the efficiency η of the tube if $\varphi_E<9.5$.

B. It explains perfectly a decrease of the power gain G.

These conclusions are very important and seem to be confirmed experimentally.

Finally, take $\varphi_E=9.5$ as an extreme limit for the operation of the modulation space. Let us calculate φ_E for the case where 10 percent of the total number of emitted electrons fall back on the cathode. The usual assumption is made that the electron flow obeys a sinusoidal law

$$n=n_0 \sin \omega t.$$

90 percent of the electrons pass before electron τ_{90} such that

$$\int_0^{\tau_{90}} \sin\theta \, d\theta = 2\times 0.9,$$

which gives $\tau_{90}=145$ degrees.

This electron, $\tau=145$ degrees, must be the last one to go through the grid; for this to happen, it must reach the grid with a velocity $v=0$; now we have $v=0$ for $\cos\theta=\cos\tau$, which gives $\theta=215$ degrees. Substituting these values for θ and τ in (10), and with $x=d$,

$$\varphi_E^2=45\times 0.17=7.65,$$
$$\varphi_E=2.76.$$

To sum up, for $0<\varphi_E<2.76$ the quasi-steady state exists. For $2.76<\varphi_E<9.5$, operation is possible in the very-high-frequency regions and for $\varphi_E>9.5$, the modulation space is useless for the production of power.

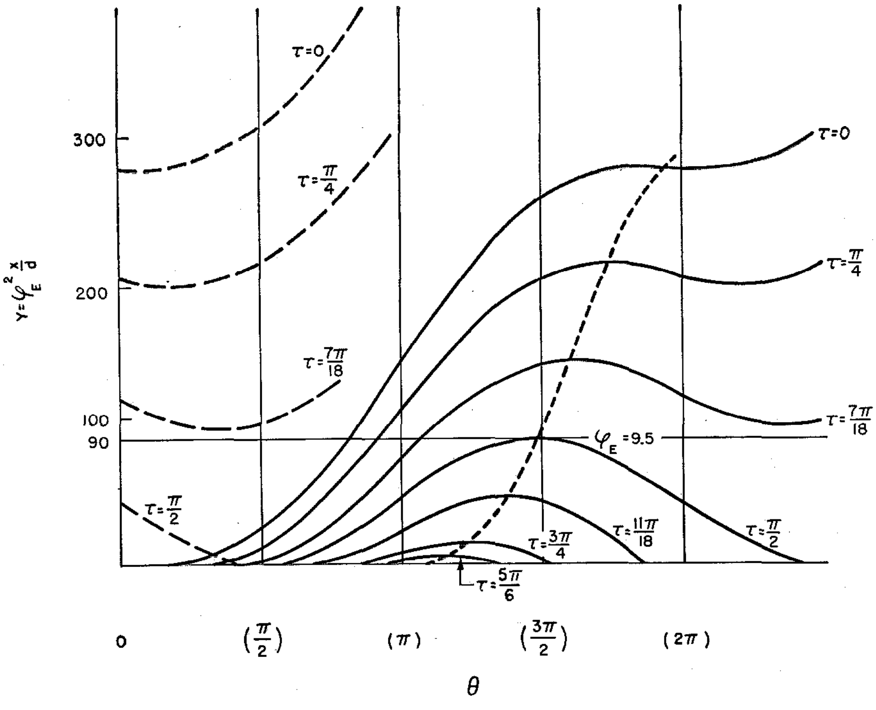

Fig. 4—Curves of y plotted against θ.

2.1.3 *Space Charges*

Llewellyn has shown that in the quasi-steady state the space charge in the neighborhood of the cathode increases by exactly 50 percent the transit time calculated with no space charge.

Thus, the phenomena remain substantially unchanged, provided we multiply the φ's by $\frac{2}{3}$. This approximation is open to criticism and we hope that a more complete theory and a graph, similar to Fig. 4 but *with* space charge, will be presented shortly. It would be most useful.

Contenting ourselves with multiplying the φ's by $\frac{2}{3}$, we finally conclude:

$$\begin{aligned} 0 < \varphi_E < 1.84 &= \text{quasi-steady state,} \\ 1.84 < \varphi_E < 6.3 &= \text{high frequencies} \\ \varphi_E > 6.3 &= \text{unusable.} \end{aligned}$$

φ_E is the particular φ adopted for the modulation space of the triode. Later on, we shall relate it to the number φ used for the whole tube.

2.2 OUTPUT SPACE

In addition to serving as an output space, this region provides room for acceleration. The field in this region results from the superposition of an unvarying field established by the accelerating source V_B and an alternating field created by the output source V_O. As V_B and V_O are of the same order of magnitude, the electron flow is a maximum at the instant the resulting potential difference $V_B + V_O$ is close to zero. Consequently, the electrons go through this space with the substantially constant velocity they acquired in the modulation space. This velocity is necessarily small, and constitutes the major drawback of the triode as an ultra-high-frequency generator.

2.2.1 *Quasi-Steady State*

In the quasi-steady state, the conditions under which ultra-high-frequency triodes are most commonly used are as follows:

A. Class-*B* operation.

B. At the time of the cycle when the electron output is maximum, the cathode is brought to a potential $-0.1\,V_B$ with respect to the grid; the grid-anode potential difference is lowered to zero (Fig. 5), which amounts to choosing $V_B = V_O$ in absolute value.

A first consequence of this potential distribution is to impose a relationship between the distances d_{FG} and d_{GA} in the tube. Referring to the article by Fay, Samuel and Shakley[8] on the effect of space charge between grid and plate, if with respect to the cathode, $V_{FG} = V_{FA}$, i.e., $V_{GA} = 0$, we cannot have

$$\frac{d_{GA}}{d_{FG}} > 2(2)^{\frac{1}{2}}.$$

If distance d_{GA} is greater than $2(2)^{\frac{1}{2}}\,d_{FG}$, a virtual cathode is formed between the grid and plate and part of the electrons are reflected to the grid. This condition is verified experimentally on triodes comprising a cathode with an equipotential surface and a practically plane structure. The grid-anode distance may be larger if we use a filamentary cathode or a cylindrical structure of small radius. In practice, in most of the ultra-high-frequency triodes $d_{GA} = 3d_{FG}$; we shall assume this condition to be fulfilled.

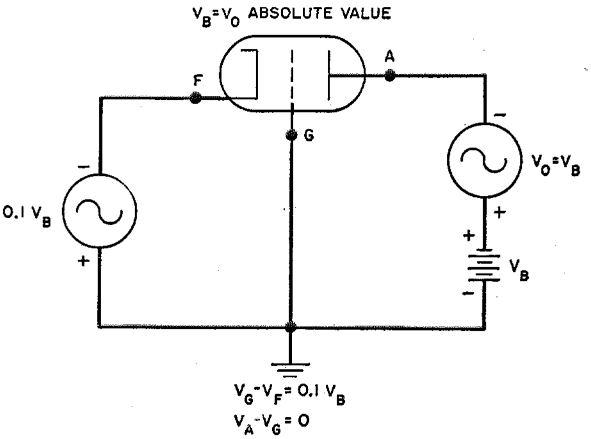

Fig. 5—Quasi-steady-state conditions under which triodes are commonly used at ultra-high frequencies.

Let us consider the values of the transit angles in the output space at peak current for these conditions.

$$\begin{aligned} V_{FG} &= V_{AF} = 0.1\,V_B. \\ d_{GA} &= 3d_{FG}. \end{aligned}$$

Assume the velocity of the electrons to be constant between the grid and plate and equal to

$$v = \left(\frac{2e}{m}\right)^{\frac{1}{2}} (0.1\,V_B)^{\frac{1}{2}}.$$

Under these conditions, the transit angle of the electrons, corresponding to the current peak in the output space, is

$$\theta_0 = 2\pi f \frac{d_{GA}}{(0.1\,V_B)^{\frac{1}{2}}} \left(\frac{m}{2e}\right)^{\frac{1}{2}}.$$

Now

$$\left(\frac{2e}{m}\right)^{\frac{1}{2}} = 6 \cdot 10^7,$$

$$d_{GA} = 3d_{FG};$$

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[8] Fay, Samuel, and Shakley, "On the Theory of Space Charge Between Parallel Plane Electrodes," *Bell System Technical Journal*, v. 17, pp. 49–79; January, 1938.

therefore:

$$\theta_0 = \frac{2\pi}{60} 3\varphi_E$$

or

$$\theta_0 = 0.3\varphi_E, \text{ approximately.}$$

It was shown that the limit of the quasi-steady-state region in the modulation space corresponds to $\varphi_E = 1.84$ or $\theta_0 = 0.56$ radian, 32 degrees, which also corresponds to the limit for the quasi-steady state in the output space. (On this subject, see Warnecke and Bernier.[9]) The fact that the limits for the quasi-steady state is reached simultaneously in the two spaces shows that the conditions of use selected are rational and explains their choice.

We can also note that

$$\varphi_0 = \frac{fd_{GA}}{V_B^{\frac{1}{2}}} = \frac{3}{10^{\frac{1}{2}}} \frac{fd_{FG}}{V_E^{\frac{1}{2}}} \neq \varphi_E$$

and $\varphi_0 = \varphi_E$ in a triode tube.

2.2.2 φ *Coefficient*

In practice, the φ coefficient of a triode has been evaluated in the following manner.

$$\varphi = \frac{fd_{FA}}{V_B^{\frac{1}{2}}}$$

where

f = frequency in megacycles
d_{FA} = cathode-anode distance in centimeters
V_B = plate volts.

This can be written

$$\varphi = \frac{f(d_{FG}+d_{GA})}{(10V_E)^{\frac{1}{2}}} = \frac{4}{10^{\frac{1}{2}}} \varphi_E,$$

$$\varphi_T = 1.26\varphi_E.$$

The limit of the quasi-steady state is thus reached for $\varphi = 1.26 \times 1.84$ or $\varphi = 2.3$.

A long experimentation has shown that it is possible, if the circuits are correctly designed, to keep the same values for η (efficiency) and G (power gain) obtained for medium high frequencies, for any value of φ up to about $\varphi = 2.5$.

This constitutes an excellent experimental proof of the above theory which, despite its lack of fine detail, has a substantial practical value.

Let us now calculate the voltage magnification factor under load Q for the output space. This plays a prominent part in television.

The reactive current through the output circuit is

$$A' = \omega VC_{GA}$$

$$= 10^{-6} \frac{fV_B}{d_{GA}}$$

$$= 1.33 \cdot 10^{-6} \frac{fV}{d_{FA}}.$$

Assuming that the total circuit capacitance is *twice* the capacitance of the active elements of the tube, if

$$\varphi = \frac{fd}{V^{\frac{1}{2}}},$$

$$A' = \varphi \cdot 1.33 \cdot 10^{-6} \frac{V_B^{\frac{3}{2}}}{d_{FA}^2}.$$

Let us now calculate the active component of the electron current. The peak current emitted by the cathode is

$$A = 2.33 \cdot 10^{-6} \frac{(0.1V_B)^{\frac{3}{2}}}{(0.25d_{FA})^2},$$

$$= 2.33 \times 16 \times 0.03 \times 10^{-6} \times \frac{V_B^{\frac{3}{2}}}{d_{FA}^2},$$

$$= \frac{V_B^{\frac{3}{2}}}{d^2} \cdot 10^{-6}, \text{ approximately.}$$

We obtain the value of the fundamental component A_{Fund} by multiplying by ¾ to account for the current collected by the grid and by the coefficient ½ for the development in a Fourier series for class B. Whence,

$$A_{\text{fund}} = 0.37 \frac{V_B^{\frac{3}{2}}}{d^2} 10^{-6}.$$

Substituting the value for A', we get

$$\frac{V_B^{\frac{3}{2}}}{d^2} = A_{\text{fund}} \times 2.66 \times 10^{-6},$$

whence

$$A' = A\varphi \times 1.33 \times 2.66,$$

whence

$$Q = 3.55\varphi.$$

Q and φ are proportional as foreseen in Section 1.4.

This relationship gives the lowest value for Q that can possibly be expected for a triode operating at normal efficiency. This relationship is valid for the quasi-steady states up to $\varphi = 2.5$ and $Q = 9$.

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[9] Warnecke and Bernier, "Contribution à la Théorie des Tubes à Modulation de Vitesse," *Revue Générale de l'Electricité;* January and February, 1942.

When using triodes with glass bulbs and circuits *outside* the tube, the total capacitance reaches four times the capacitance of the active elements, which doubles the above value and gives $Q = 7.1\varphi$, whence $Q = 18$ for $\varphi = 2.5$.

This is a fundamental relationship in the technique of modern wide-band television transmitters operating at very high frequencies.

2.2.3 *Transit Angles Not Negligible*

Cases where transit angles are no longer negligible correspond to values of φ_E between 1.84 and 6.3; φ then varies from 2.32 to 8, and the transit angle in the output space varies from 32 to 110 degrees. This value of 110 degrees, which was calculated without taking into account the space charge, is a high value, but it is still lower than 180 degrees for which the efficiency of the output space is still appreciable.

This shows that the amplification of a triode G decreases much more rapidly than the efficiency. This has also been confirmed by experiment.

Of course, when $G = 1$, interest in the tube as an amplifier vanishes completely and the power it can deliver as an oscillator drops to zero. We do not know exactly the value of φ corresponding to such a case. For $\varphi = 8$, certain triode oscillators still have an efficiency of 5 to 10 percent, which is outside the field in which we are interested. Table I shows relations of φ, Q, and general operating conditions;

$$\varphi = \frac{fd}{V^{\frac{1}{2}}} \quad \text{and} \quad Q = \frac{f}{\Delta f},$$

where

f = frequency in megacycles,
d = filament-anode distance in centimeters,
V = plate volts,
Δf = total bandwidth in megacycles.

TABLE I

Operation of Triodes at Very High Frequencies

φ	Q	Remarks
0 2.5	0 18	Quasi-steady state. Results similar to those obtained at low frequencies.
2.5 8	18 above 60	Important transit times. Power gain drops rapidly. Output drops slowly.

A complete study of the dimensioning of triodes would be beyond the scope of the present study.

It will be realized easily that it is possible to determine completely all the dimensions and operating conditions for the optimum case of a triode oscillating at a frequency f.

If we are dealing with a pulsed tube or if f is above 600 megacycles, the maximum emission is the dominant factor in computing the dimensions.

If we are dealing with a continuous-wave tube operating at a frequency lower than 600 megacycles, the maximum dissipation per unit surface is the dominant factor. It is then possible to calculate for each frequency the most favorable distances between electrodes, plate voltage V_B, cathode emission A, pass band Δf, etc.

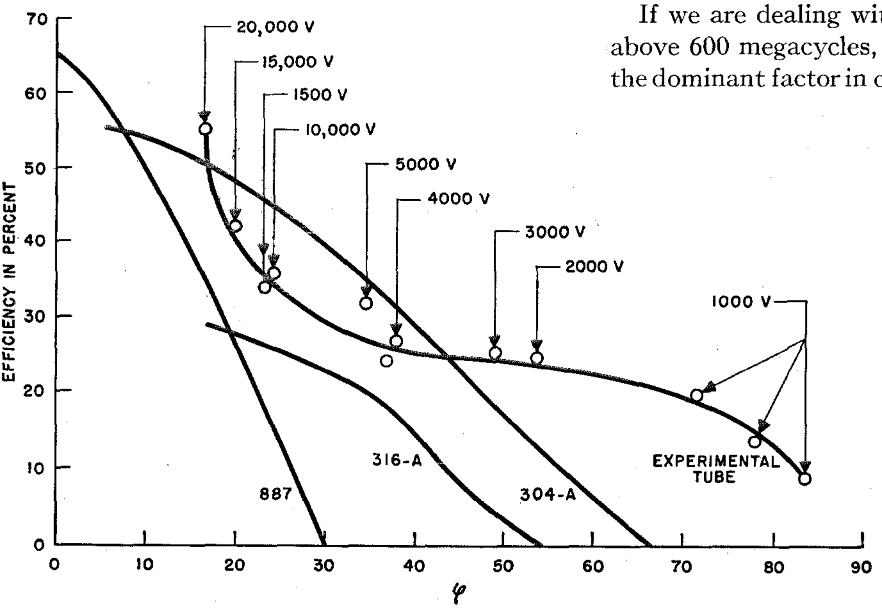

Fig. 6—Efficiency plotted against φ for three prewar triodes and an experimental tube.

The total area of the tube is not controlled by the above calculations, and may be adjusted to the total power required; its design is influenced greatly by the mechanical conditions of the construction and by the stiffness of the elements.

Fig. 6 shows curves of efficiency as an oscillator of three prewar types and one experimental tube. It is interesting to note the accuracy with which the points relative to the latter tube are located on a single curve, despite the extreme values of the voltages used.

3. *Conclusion*

It is a known fact that it is often difficult to determine whether unsatisfactory operation is attributable to the tube or the circuit. Dimensional analysis has contributed to the solution of this problem, and despite the lack of refinement of the theory and the unsubstantial nature of certain hypotheses, the results are far from being useless.

New types of triodes[10] have been developed, which are useful in the construction of high-definition color-television transmitters[11] delivering 1 kilowatt on a wavelength of 0.6 meter with a modulation band extending to 10 megacycles, and of frequency-modulated broadcast transmitters producing 100 kilowatts at 3 meters. To obtain such results, circuits of a new type must be used.

4. *Acknowledgment*

The use of dimensional analysis for the study of tubes has been the object of unpublished memoranda by Messrs. Goudet and Boggs. I also want to thank for their assistance in this work, my colleagues at Federal Telecommunication Laboratories and, more particularly, Messrs. Emile Labin and A. R. Vallarino.

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Flashing Signal for Railway Crossings

By V. C. MEEUWS

Bell Telephone Manufacturing Company, Antwerp, Belgium

RAILWAY CROSSINGS in Belgium, as in many other European countries, are commonly protected by gates operated by signalmen. The considerable increase in automobile and other road traffic and the expansion of the highway network have made desirable the replacement of the manually operated gates by signals that are controlled automatically.

Red and green lights are employed to control the traffic over highways that cross the railroad. In addition, an audible warning is given by an electric gong when a train approaches. The red light flashes at a rate of 80 times per minute and the green light flashes at 40 times per minute.

1. *Flashing Mechanisms*

Various types of circuit interrupters have been developed for flashing these lamps. The service is rather heavy and where mechanical interruption is used, trouble with electrical contacts has been encountered. The failure of the interrupter would bring about a hazardous condition and serious accidents might easily result. In addition, maintenance costs of such equipment are not insignificant. Consequently, the problem of flashing these lamps by nonmechanical means has been given careful attention.

The problem, in general, is to flash two, three, or four signal lamps, each of 60 watts rating, at 80 or 40 pulses per minute. The lamps are normally operated from the 130-volt alternating-current supply. If one or more lamps go out of service, the remaining lamps must continue to flash. The "On" and "Off" times are to be approximately equal. The control circuit should, so far as possible, be common to both the red and green lamps.

2. *Electric Flasher*

In Fig. 1, the current flowing through the lamp is in part controlled by the inductance of the two windings, *W*1 and *W*4. When the power is applied to the circuit, the inductance, and hence the impedance, of *W*1 is so high that the lamp fails to light. Rectifier 1, which is connected across the power supply, applies a direct voltage to the capacitor *C* through *R*1 and *W*2. As the capacitor charges, the voltage across it increases until it is high enough to ignite the gaseous discharge lamp shunted across *C* and *W*2. This discharges the capacitor until the voltage across it is too low to maintain the gaseous discharge. The cycle then repeats and its frequency may be controlled by varying either *C* or *R*1. This is the well-known relaxation oscillator.

Only a small amount of energy is available from the relaxation oscillator to produce a magnetic field in the iron core on which the three windings are mounted. The function of the magnetic field is to reduce the inductance and impedance of the winding *W*1 to allow an increased current to flow through the signal lamp. Fortunately, the energy required is not directly related to that taken by the signal lamps but may be very much less. The frequency at which the oscillations occur is significant as more energy is required to change the magnetic field as the frequency is increased. This is a limitation on the control possible with a given gaseous discharge tube. To stabilize the control and increase the effective power range of operation, use is made of the two windings, *W*3 and *W*4, and rectifier 2.

When *C* discharges through *W*2 and the gaseous tube, the inductance of *W*1 decreases and the current through *W*4 and the lamp increases. This increases the voltage drop across *W*4 and the direct current through *W*3. *W*3 is arranged to provide a magnetic field in the same direction as that from *W*2 under these conditions. Consequently, the inductance of *W*1 is further reduced. An increased current through *W*4 and the lamp results and this process continues until magnetic saturation of the core prevents any further reduction in the inductance of *W*1.

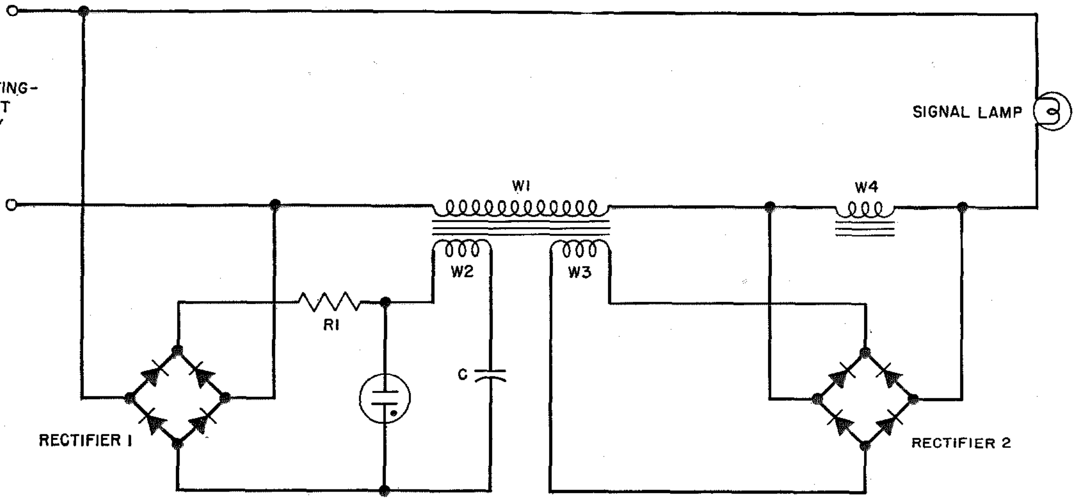


Fig. 1—Circuit used for flashing signal lamps. A relaxation oscillator employing a gaseous conduction tube controls the rate at which the lamp is flashed.

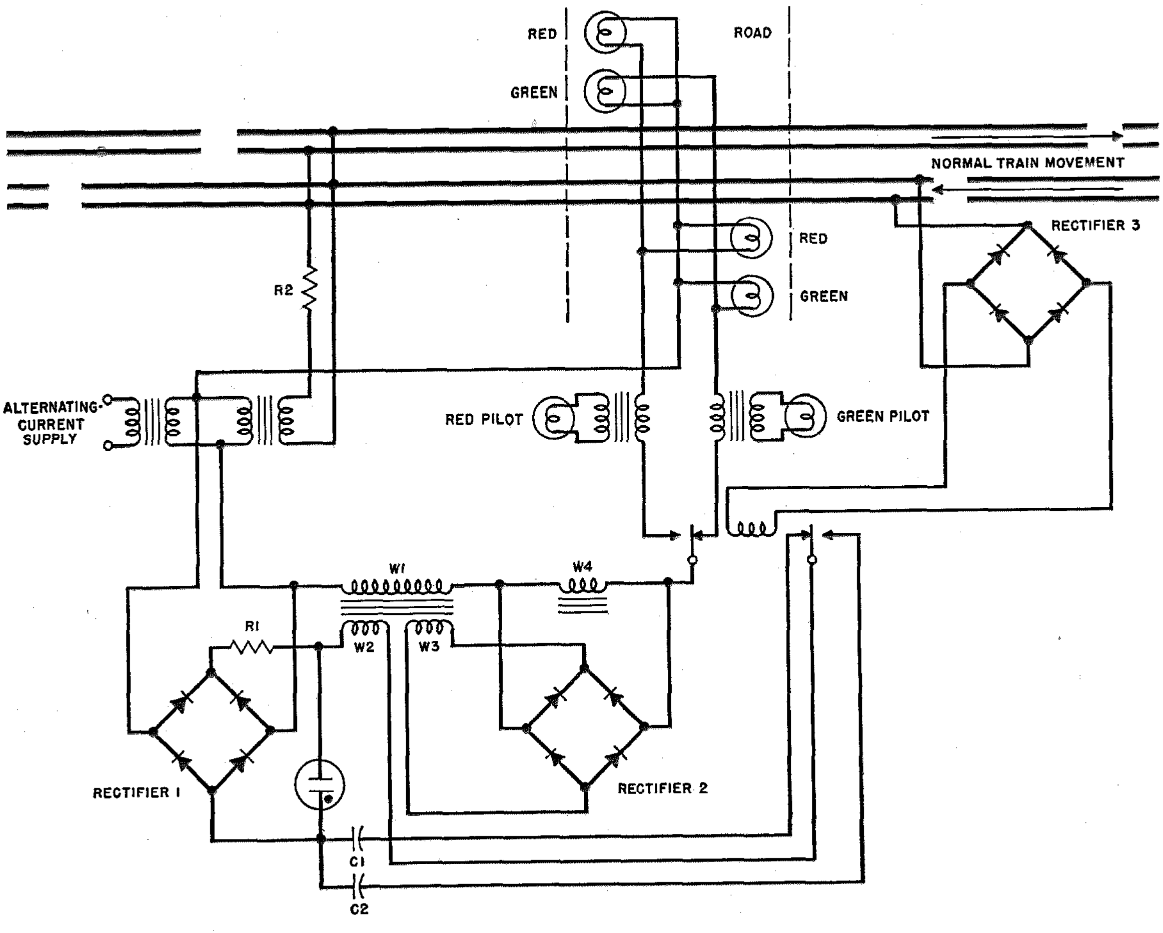


Fig. 2—Arrangement used at a typical railway crossing. The railway tracks are indicated by the heavy horizontal lines at the top and the dashed lines designate the highway for automobile and other traffic.

When the capacitor C discharges to a voltage at which the gaseous tube is no longer conducting, a current from the rectifier flows through $R1$ and $W2$ to recharge the capacitor. This current through $W2$ is in the opposite direction to that which resulted from the discharge of the capacitor. Thus it produces a magnetic field which opposes that produced by the current flowing in $W3$. This increases the inductance of $W1$, which reduces the current through $W4$ and the lamp. The reduced current through $W4$ decreases the rectifier current through $W3$, further increasing the inductance of $W1$. This process continues until the inductance of $W1$ is sufficiently high to extinguish the lamp.

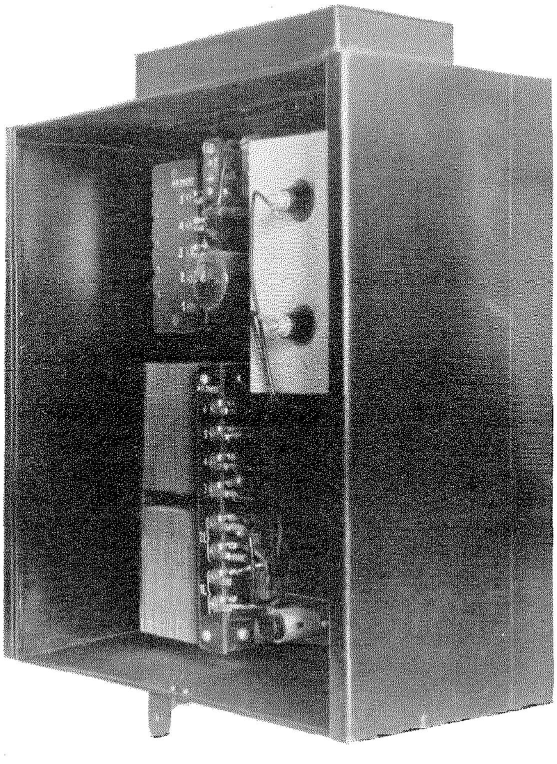


Fig. 3—The electronic flashing control equipment is mounted in a metallic case.

3. Railroad Crossing Installation

The arrangement used in a typical railroad crossing installation is shown in Fig. 2. Red and green signal lamps are located at each side of the railway facing the direction in which the traffic on the highway approaches the railroad.

At a suitable distance on each side of the crossing, the rails are insulated electrically from the rest of the railway system. These sections which are crossed by the highway are supplied with power at a voltage sufficiently low to offer no hazard to anyone making any contact with the rails. From these rails, rectifier 3 supplies power to a relay which under normal conditions supplies energy to the green signal lamps. When a train enters the protected region of either track, the input to the rectifier is short-circuited and the relay disconnects power from the green lights and applies it to the red signals. A resistor protects the transformer from burning out during this condition.

The pilot lamps are installed in the cabin of the nearest attendant to permit supervision of the operation of the signal lamps. The primary of the transformer to which the pilot light is connected is adjusted in accordance with the number of lamps that are being flashed.

A second set of contacts on the relay inserts a 40-microfarad capacitor in the relaxation oscillator circuit to flash the red lamp at a rate of 80 times per minute and an 80-microfarad capacitor to flash the green lamp at 40 times per minute.

The electronic flasher equipment is mounted in a metallic case 270 millimeters (10¾ inches) square by 157 millimeters (6¼ inches) deep (Fig. 3).

Although originally developed for the Société Nationale de Chemins de Fer Belges, its use is not limited to railway crossings but it may be applied to other purposes. It is also being used for street-traffic control.

In Memoriam

HENRY MARK PEASE

Henry Mark Pease died on March 7, 1947, in New York after a brief illness. Born in Malta, Illinois, on December 19, 1875, he was a member of the ninth generation of the Pease family in the U. S. A.

Mr. Pease joined the Western Electric Company in Chicago in 1898, shortly after his graduation from the University of Illinois with a Bachelor of Science degree. After serving in the wiring shop, he was assigned to installation of central-office equipment in various American cities, and to supervision of installation in Cincinnati. Later, he returned to Chicago as an equipment engineer.

In December, 1902, Mr. Pease was transferred to the London branch of Western Electric to supervise the installation of central-battery telephone exchanges for the British Post Office and the National Telephone Company. After serving as chief engineer, he was transferred in 1907 to the sales department, becoming sales manager in 1909, and assistant manager of the company in 1913. In 1918, he was made managing director.

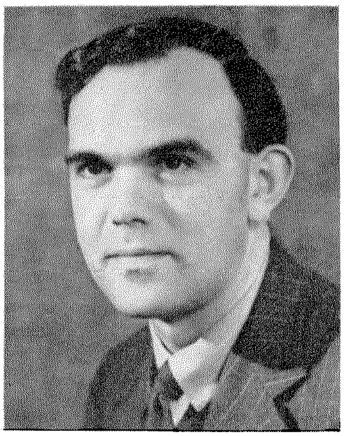
From 1908 to 1922, Mr. Pease also had charge of the extension of loaded-cable telephone transmission systems in Great Britain, Holland, Sweden, and Italy. In London, he organized the long-distance cable-laying department.

In 1922, he took an active part in forming the British Broadcasting Company, becoming one of its original directors. International Western Electric, London, installed one of the first broadcasting stations in England. The following year, Mr. Pease negotiated a contract with the British Post Office for the first transatlantic radiotelephone transmitting station.

In 1925, International Telephone and Telegraph Corporation purchased International Western Electric Company and changed its name to International Standard Electric Corporation. Concurrently, the name of Western Electric, London, was changed to Standard Telephones and Cables, Limited. Mr. Pease continued as managing director of Standard Telephones and Cables until 1928 when he was appointed European general manager of International Standard Electric Corporation.

Mr. Pease returned to New York in 1933 as vice president and a director of the International Standard Electric Corporation. In 1941, he became president and last year assumed the post of first vice-chairman of the board. In 1937, he was elected vice president and a director of International Telephone and Telegraph Corporation. In addition, Mr. Pease served as vice president and a director of Federal Telephone and Radio Corporation, and as a director of both International Telecommunication Laboratories, Inc. and International Telephone and Telegraph Corporation, Sud America.

Contributors to This Issue



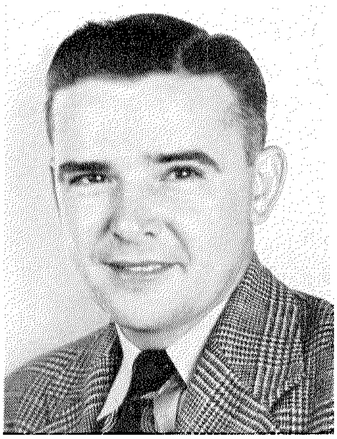
H. J. Barker

H. J. Barker was born at Halifax, Yorkshire, England, on April 14, 1915. He obtained an Honours degree in 1937 in physics from the Royal College of Science, London. He then did postgraduate work in electrical communications at City and Guilds College.

Since 1938, he has been with Standard Telephones and Cables, Limited, working in the carrier systems design and planning group at North Woolwich.

• • •

Vernon F. Clifford was born on September 30, 1915, at Greenport, New York. He graduated from Brooklyn Technical High School.



Vernon F. Clifford

After engaging in audio-frequency development work on motion-picture and sound-distribution systems for Transformer Corporation of America, he joined the direction-finder group of International Telephone and Radio Laboratories in 1941. He has worked on marine, mobile, aircraft, and fixed installations for the armed forces and for civilian use. At present, he is an engineer in the direction-finder division of Federal Telecommunication Laboratories.

• • •

Milton Dishal was born on March 20, 1918, in Philadelphia, Pennsylvania. Temple University conferred on him two degrees, the B.S. in 1939 and M.A. in 1941. He was a Teaching Fellow in physics in that University from 1939 to 1941.

Mr. Dishal entered Federal Telecommunication Laboratories in 1941 and is now a senior engineer in the development of radio receivers having special characteristics.

Mr. Dishal is a Senior Member of the Institute of Radio Engineers.

• • •

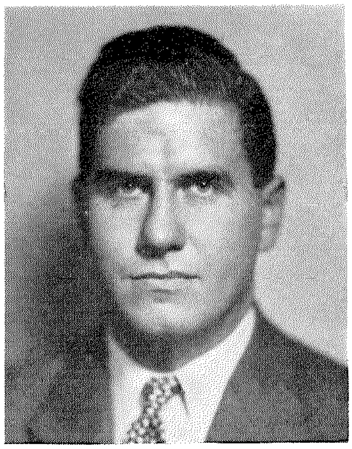
John Kemp received his engineering training at the Imperial College of Science and Technology, London, receiving the diploma in 1921.

On completing college, he joined the International Western Electric Corporation. After a course of study in America, he specialized in the engineering and manufacture of toll cables. At the outbreak of the war, he was assigned to theoretical studies and is the author of several papers on wave guides.

He was elected a Member of the Institution of Electrical Engineers in 1935.

• • •

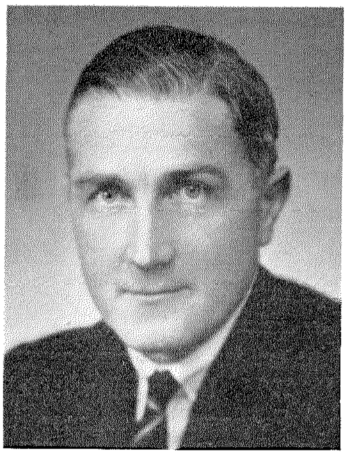
William Kidd was born in Newcastle-on-Tyne, England, and was educated in mechanical and electrical engineering at Rutherford College and Durham College of Science.



Milton Dishal

He served an apprenticeship in mechanical and marine engineering and obtained some commercial experience with the Yorkshire Brass and Copper Company. He was then employed as a draftsman at Palmers Steel and Iron Company. Later he became a draftsman and assistant to the works manager of the North Eastern Marine Engineering Company.

After a short period as a draftsman with the Manchester Corporation Electricity Department, he became assistant engineer at the Stuart Street generating station, then assistant to the deputy chief engineer, and finally chief construction engineer.



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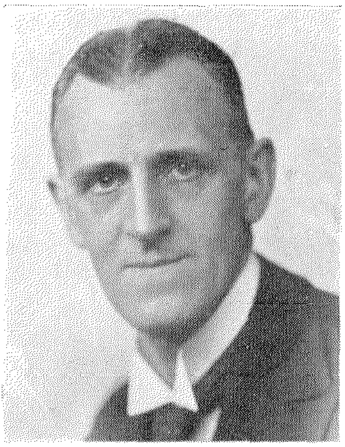
John Kemp



William Kidd

Mr. Kidd is a Member of both the Institution of Mechanical Engineers and the Institution of Electrical Engineers. He has served on the Council of the Institution of Electrical Engineers, as chairman of the North-Western Centre and of the North-Western Transmission Group, and on the North-Western Committee. He is a member of the British Electrical Research Association Transformer Committee.

Mr. Kidd received from the Institution of Electrical Engineers the "Institution" premium in 1934 for his paper on "Automatic Voltage Control" and, jointly with E. M. S. McWhirter, the "John Snell" premium in 1945 for the paper "Operational Control of Electricity Supply Systems" reprinted in this issue of *Electrical Communication.*

• • •

Gerard J. Lehmann

Gerard J. Lehmann was born in Paris, France, on April 6, 1909. He received an engineering degree from Ecole Centrale in 1931.

On leaving school, he entered the employ of Sadir, becoming technical director in 1939.

In 1940, after leaving the French Army, he became a member of the Lyon laboratory staff of Le Matériel Téléphonique. He was transferred to Federal Telephone and Radio Laboratories in New York in 1943 and returned to France in 1945.

In addition to research work at Laboratoire Central de Télécommunications, Mr. Lehmann has been teaching at Ecole Centrale and in 1942 was appointed professor of direction finding and radio navigation at Ecole Supérieure d'Electricité.

Mr. Lehmann is a Senior Member of the Institute of Radio Engineers.

• • •

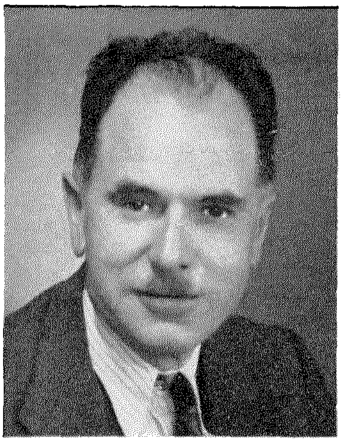
Victor Constant Meeuws was born at Antwerp (Berchem) on August 12, 1891. He was engaged by the Bell Telephone Manufacturing Company in 1913 and worked as a draftsman on the first rotary equipment manufactured by the company. After serving in the army from 1914 to 1919, he was placed in charge of the wiring section. In 1930, he was assigned to equipment design of the 7-D Rotary System. Four years later, he assumed the duties of power engineer.

In 1933, Mr. Meeuws was granted the title of Technical Engineer by a committee set up by law.

• • •

Carlos Pelaez Perez Gamoneda joined Compañía Telefónica Nacional de España in 1924, having worked for some years previously with power supply companies and the Spanish military communications department. He served also as a professor in an engineering school in Spain. He has also worked on modulation transformers and high-powered radio equipment in Le Matériel Teléphonique in Paris.

In Compañía Telefónica Nacional de España, he was director of instruction, assistant chief engineer, and then chief engineer of the radio services. During the civil war in Spain, he was stationed in Madrid as subdirector general and also as vice president of International Telephone and Telegraph Corporation (España).



Victor C. Meeuws

Mr. Pelaez is now in charge of the technical department of Standard Electrica in Lisbon and is assistant technical director of Standard Electrica S.A., Madrid, Compañía Radio Aerea Maritima Española and Sociedade Radio Argentina of Spain.

• • •

D. P. J. Retief was born in Pretoria, South Africa, on November 5, 1904. He received the B.Sc. (Eng) and B.A. degrees from the University of Capetown in 1929.

From 1930 to 1932, he was with Siemens in Berlin. He then returned to South Africa and joined the

Carlos Pelaez

D. P. J. Retief

Department of Posts and Telegraphs. He visited England in 1938 and 1944. As transmission engineer, he is now in charge of the development of long-distance communication.

Mr. Retief is an Associate Member of the Institution of Electrical Engineers and a Senior Member of the Engineers Association (South Africa).

• • •

George Taylor Royden was born at Fort Clark, an army post in Texas, on June 20, 1895. He received a B.A. degree in 1917 and an engineering degree in 1924 from Stanford University.

He was employed by Federal Telegraph Company at Palo Alto, California, on part time in 1916 and full time after graduation from college. He was engaged in the design of arc transmitters of powers up to 1000 kilowatts.

From 1919 until 1925 he was at Mare Island Navy Yard. His duties included work on Navy radio stations in San Diego, Hawaii, and Alaska.

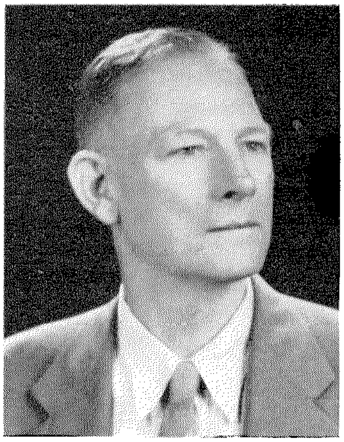
He returned to Federal Telegraph Company in 1925 to do research work on broadcast receivers for operation on alternating current.

In 1927, he joined the newly organized Mackay Radio and Telegraph Company becoming division engineer. From 1936 to 1946, he was with Federal Telegraph Company in Newark, New Jersey. He then returned to Mackay Radio and Telegraph Company.

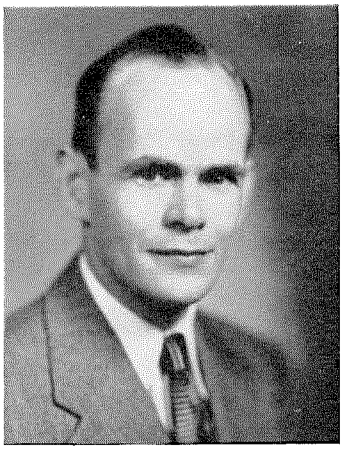
Mr. Royden is a Fellow of the Institute of Radio Engineers and a Member of the American Institute of Electrical Engineers and of Sigma Xi.

• • •

Norman H. Young was born in Philadelphia, Pennsylvania, in 1913. He received the B.S. degree in electrical engineering from Pennsylvania State College in 1934 and the M.S. degree in 1935.



George T. Royden



Norman H. Young

From 1935 to 1942, he was engaged in television engineering for the Philco Corporation and had charge of the transmitter of television station WPTZ.

In 1942 he became a department head in Federal Telecommunication Laboratories and during the war was largely concerned with the application of pulse-time modulation to military communication equipment. At the termination of the war, he was responsible for the engineering of the color-television transmitter for the Columbia Broadcasting System. He has done additional work on receivers and studio equipment for color television.

Mr. Young is a member of Eta Kappa Nu and a Senior Member of the Institute of Radio Engineers.

• • •

For biographies and photographs of Trevor H. Clark and D. D. Grieg, see Volume 24, Number 2, pages 275 and 277.

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All America Cables and Radio, Inc., New York, New York[4]

The Cuban All America Cables, Incorporated, Havana, Cuba[2]

Sociedad Anónima Radio Argentina, Buenos Aires, Argentina[5]

[2] Cable service. [3] International and Marine Radiotelegraph services.
[4] Cable and Radiotelegraph services. [5] Radiotelegraph service.

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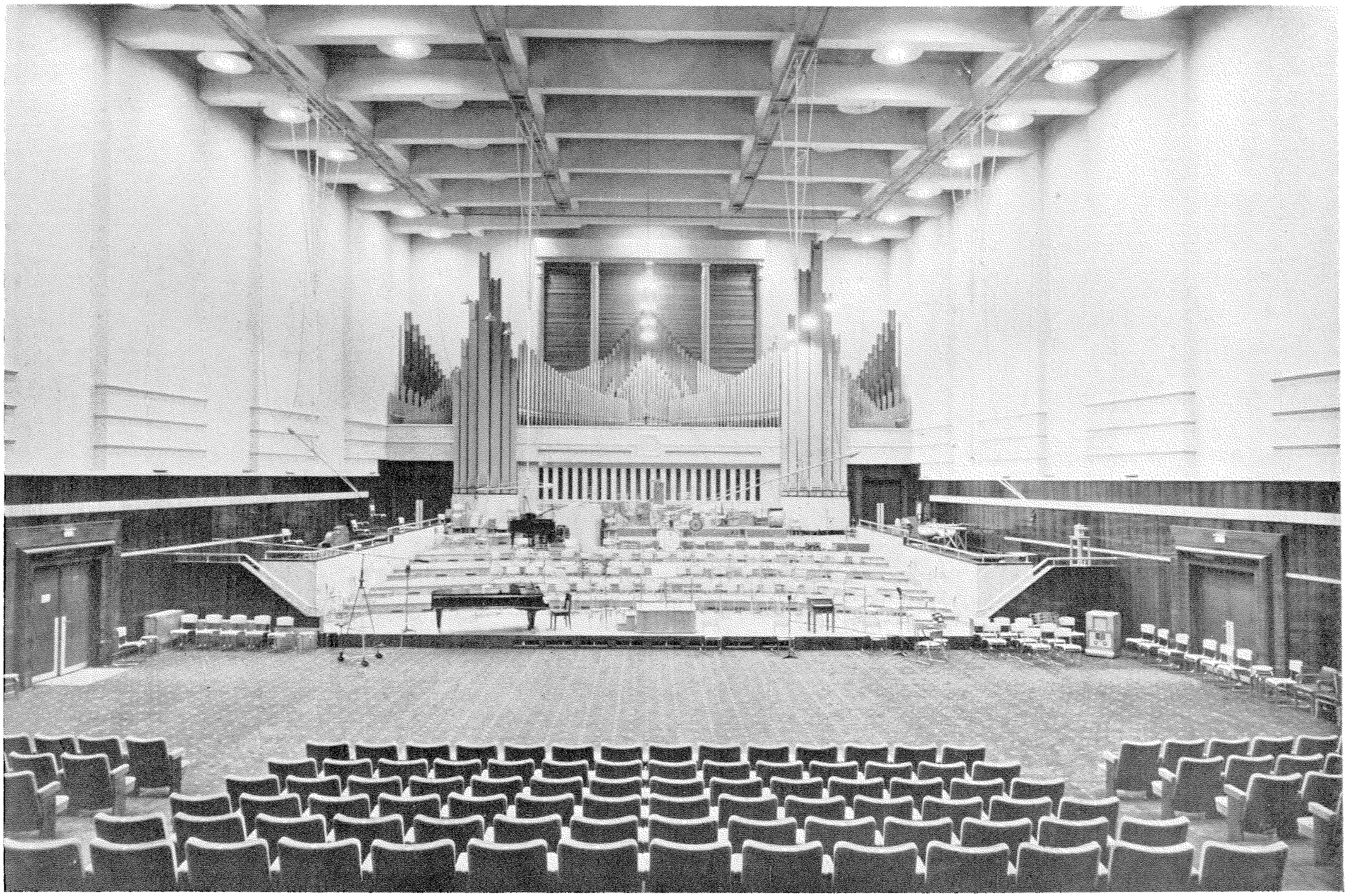


The great concert hall in Broadcasting House of the Institut National Belge de Radiodiffusion in Brussels, Belgium.

Speech-Input Equipment of Brussels Broadcasting House

By FERNAND MORTIAUX

Institut National Belge de Radiodiffusion, Brussels, Belgium

SHORTLY before the start of World War II, the broadcasting headquarters of the Institut National Belge de Radiodiffusion were completed in Brussels. A new building, shown in Fig. 1, provides for 19 studios, 5 recording rooms, and all of the communication, signaling, and supervisory accessories essential to a modern broadcasting system. A view of the concert hall is shown in the frontispiece.

This installation provides for two simultaneous main programs and one colonial program; each main channel supplies about 4500 hours of broadcasting time per year. A studio includes an associated announcing room and a control room from which all studio activity is visible. All announcements are made audible in the studio and all studio material is reproduced in the announcing room. The control room personnel, of course, hear everything.

Each channel, which comprises all equipment from the microphone or pickup to the outgoing line to the radio transmitter, is flat within ±2 decibels over a frequency band from 30 to 14,000 cycles per second, has less than 1-percent harmonic distortion, and has minimum phase distortion. Noise of all types, including that from vacuum tubes, resistance thermal agitation, and disturbances from alternating-current supply circuits as a result of inadequate filtering or induction, is at least 80 decibels below a reference level of 1.55 volts in a 40-ohm circuit, which corresponds to 100-percent modulation of the radio transmitter.

The number of studios is determined not only by the programs being broadcast but by the time needed for rehearsals. This usually amounts to three or four times the actual broadcasting requirement and, to provide two national and one colonial program, the studio facilities are often engaged up to 80 percent of their maximum possible utilization.

Smaller installations have found it convenient to centralize all technical equipment in a single room. Such an arrangement would have led to considerable complexity, and continuity of operation, flexibility, and supervision would have been extremely difficult. The decentralized system adopted groups the amplifiers and control equipment near the studio with which they are associated. The switching means for connecting the various studio outputs to the radio transmitter, recorders, or other terminals, are concentrated in separate rooms to which only the broadcasting supervisors have access. This decentralization introduces a certain amount of automatic switching to enable the required connections to be set up without damaging delay.

A studio, together with its announcing and control rooms and all necessary technical equipment to supply an outgoing signal at the reference level corresponding to full modulation of the radio transmitter, is designated as an "alpha." It is a complete source of program material. Alphas designed for speech only do not have announcing rooms and those for reproducing recorded material have no studios. An alpha is under the supervision of a single operator who is responsible for its technical performance, including the outgoing volume level. There are 24 alphas in the installation.

A "lambda" is a destination that receives the outgoing signal from an alpha. The 10 lambdas consist of three lines to radio transmitters, five recording equipments, and two general mixing desks. The lambdas are controlled by the broadcast supervisor who is responsible for technical quality, volume control, and continuity of transmission. Switching means are provided whereby each alpha may be connected to any one lambda.

All incoming and outgoing telephone lines terminate in a room called the "lines alpha." All outside programs, including relays of foreign broadcasts, are distributed from the lines alpha. Similarly, programs from within the building, intended for retransmission abroad, leave from this room. All outgoing lines from alphas and lambdas are brought out on jacks on a central switchboard. Lines alpha contains a manual telephone exchange as well.

The automatic program distribution service permits any "subscriber" inside the building to listen to any alpha by plugging in headphones or an amplifier and loudspeaker and dialing the designated number.

Fig. 1—Broadcasting House of Institut National Belge de Radiodiffusion in Brussels.

1. Alpha

A studio alpha is a basic source of program material and consists of a studio and its associated announcing and control rooms. It may obtain program material from four different sources: the studio, the announcing room, a circuit from one of the recording rooms, and a transmission from outside the alpha. This fourth source may be the retransmission of a foreign program, a program originating outside the building (outside broadcast), material transmitted from any other alpha, or a time signal. These four sources are connected to preamplifiers, mixed, and then passed through a second amplifier, which is called the line amplifier, for transmission to the lambdas at reference level.

A technician and an operating musician who has charge of microphone placement and general acoustic balance of the program are located in the control room. They may listen to the program either from a loudspeaker or headphones. The producer and announcer will be in the announcing room and will hear the program from a loudspeaker. Automatic switching is provided to silence a speaker when the microphone in the same room is in service. The control desk is illustrated in Fig. 2.

In monitoring, the volume variation at the microphone, which may reach 80 decibels or so, is reduced to approximately 40 decibels for transmission to the lambdas.

The Comité Consultatif International Téléphonique recommends a limit of 1.55 root-mean-square volts for transmitting program material over a telephone line. This value has been adopted as reference level and will modulate the radio transmitters fully.

The minimum voltage is governed by the noise generated in the transmitter and is about 60 decibels below the 100-percent-modulation level. The noise level from the alpha is reduced to more than 80 decibels below reference level. Modern amplifiers permit this low noise level even with alternating-current operation of vacuum-tube heaters. The value of this wide volume range is evident when it is remembered that a large symphonic orchestra may produce a sound level varying over a range of 60 to 100 decibels.

The characteristic impedance of all program transmission lines was standardized at 40 ohms. This low impedance makes the line quite susceptible to induced magnetic disturbances but reduces the effects of capacitance, thus permitting a wide frequency band to be transmitted on some rather long circuits within the large building.

1.1 Volume Controls

The microphone volume controls are of the symmetrical series type to minimize losses. Because of the small electroacoustic efficiency of the microphones, the load resistance may drop below the internal resistance without producing harmonic distortion. This control has an attenuation of 2 decibels for each of 30 steps and a maximum of 110 decibels in the "out" position.

The mixing operation for the studio microphones is the only one done at low level. Experience had shown that contact and induced noises would not be damaging if the microphone lines were reasonably short and were separated effectively from power lines. Low-level mixing avoids the use of an amplifier for each microphone, thus simplifying operation and maintenance.

The amplifier volume controls are of the *L* type and are connected in series to permit fading in and out of programs from different sources, such as studio and announcing microphones and reproducing pickups. The load resistance at the output of the amplifier is higher than its output impedance. The attenuation steps are the same as for the microphone control.

The third volume control in the channel is the master or main control. It is of the ladder type, but the resistance normally connected in series with the rotating arm has been omitted. Although this affects the constancy of impedance with attenuation, it reduces the insertion loss to a minimum. Starting with the minimum attenuation position, the first 25 steps are of 2 decibels each, the next four steps give, progressively, 53, 58, 67, and 84 decibels, and the off position, 100 decibels of attenuation.

1.2 Recorded Material

Turntables and pickups for reproducing recorded material are located in the control room and may be seen in Fig. 3. The output of the pickup passes through a 40-ohm volume control of the *T* type and an equalizing network. Low-pass filters cutting off at either 4800 or 6000 cycles may be switched into the circuit for noise suppression. These program sources are particularly useful for sound effects and incidental music in plays and, of course, for completely recorded musical programs.

A synchronous motor operating at 1000 revolutions per minute drives a turntable weighing approximately 11 kilograms (24 pounds) through a speed reducer and a planet gear. An oil buffer is located immediately beneath the turntable to dampen vibrations from the drive system. The

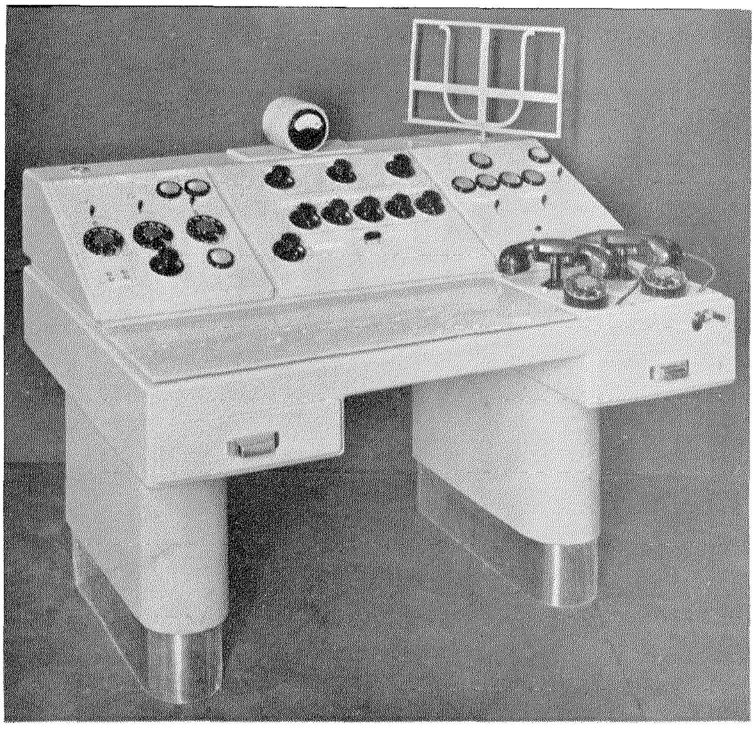


Fig. 2—Alpha control desk.

instantaneous variations in turntable speed are less than 0.5 percent. It is possible to shift the relative position of the turntable and spindle under operating conditions to synchronize two records and obtain a perfectly smooth transition from one to the other. Thus, lengthy orchestral selections can be transmitted without interruption.

The pickup arm has large inertia and is balanced by a counterweight. An indexing device enables the needle to be set in any predetermined groove on the record.

1.3 TALK-BACK

The microphones in either the control or announcing rooms may be connected to a loudspeaker in the studio, the necessary switching being provided to disconnect the studio microphone. This permits the studio personnel to hear the announcements.

The use of the microphone in the control room permits instructions to be given to the performers in the studio, which is particularly valuable during rehearsals.

In conjunction with the automatic program distribution service, talk-back from elsewhere in the building is a particularly important facility when several studios are involved in a single broadcast.

1.4 SIGNALING

Signal lights are installed in the different units of an alpha to facilitate operation and reduce the possibilities of errors. The fading-in of a studio microphone automatically energizes the corresponding signal lights through auxiliary contacts on the volume control. These contacts also control those loudspeakers that must be silenced to avoid singing. Special keys are provided to disable these auxiliary controls under special circumstances. To keep a permanent check on the more important signal lamps, a small pilot lamp, mounted near the control, is in series with the main lamp.

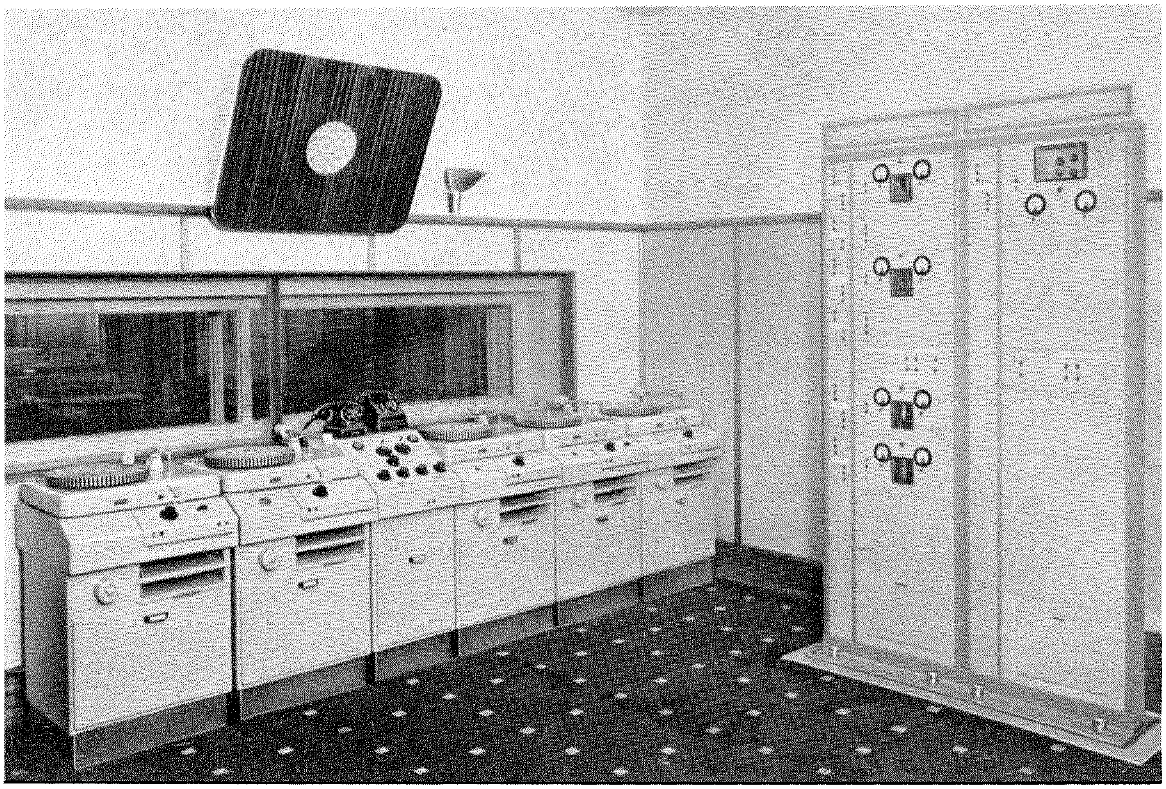


Fig. 3—Turntable desks and equipment racks in an alpha control room.

1.5 ALPHA-LAMBDA SIGNALING AND INTERCONNECTION

An automatic switching system, normally referred to as "antenna selection" or "lambda selection," is provided to permit rapid connection of any alpha to any lambda.

The alpha technician depresses a key, which sets up a connection to the antenna-selection equipment and operates an indicator lamp. A standard telephone dial is then manipulated to send impulses to a five-level selector of the step-by-step type, individual to each alpha. At the end of this operation, the selector is in the position corresponding to the required lambda. A green light then appears on the control desk of the alpha and on the corresponding panel of the lambda.

If the desired lambda is free, the branching relays operate immediately and this is indicated by red lights in the alpha and lambda. If, however, the lambda is already engaged by an alpha, a vertical locking device on the switchboard prevents the branching of the new alpha. This is a waiting position and is indicated by a green light in the alpha and lambda. A third condition is when the lambda is engaged and another alpha is in the waiting position. Then, a new alpha and all following alphas will receive a flashing green signal.

In this latter case, the broadcast supervisor in the lambda can intervene and by means of a release key any alpha may be disconnected temporarily. It is very exceptional to have more than two alphas in the waiting position.

To inform the technician of an alpha that his demand for a particular lambda has been correctly transmitted, the step-by-step selector switch is connected to the program multiple of the program distribution system. Thus, dialing of a desired lambda operates a selector in the program distribution room and lights a white signal on the control desk in the alpha when the latter is connected to the lambda. This indicates unmistakably that the antenna selection has been correctly made and transmitted.

To disconnect an alpha from a lambda, the alpha technician returns the "antenna selection" key to its original position. A waiting alpha is then automatically connected to the lambda and its technician is informed by the green signal changing to red.

1.6 CABLING

For transmission purposes, each alpha is linked to 10 pairs of wires. One pair is the normal outgoing line, a second pair operates a program meter, three trunk pairs go to the lines alpha, four pairs are for the automatic program distribution service, and the final pair is for talkback. The signaling circuits are contained in a cable consisting of one quad and 15 pairs. These cables build up what is called the "multiple" of the alphas in the lambdas and terminate in the lines alpha.

2. Lambda

The lambdas have branch relays for connection to the alphas. Equipment is also provided to permit the broadcast supervisor to monitor the program being transmitted.

The channel from an alpha terminates on a switchboard in each of the lambdas. A second line, connected to the output of the program meter of the alpha, also terminates on the general switchboard of the lambdas. This results in a large switchboard of the crossbar type, connections being made at the intersections of horizontal lines representing the alphas and vertical lines representing the lambdas. Contact is established by two special branching relays which are operated automatically by a call from an alpha if the lambda is free.

The flexibility of automatic branching was considered necessary to insure rapid switching to a series of different alphas which might be required for the transmission of certain programs by a given lambda.

As a precaution against failure, the single relay in the transmission channel proper may be bridged by hand.

To avoid the generation of noise, each contact spring of the branching relays is equipped with two independent contact points in parallel. In addition, these contacts are "wetted" by a special circuit arrangement that automatically sends a very weak direct current through the contacts as they close. This current is gradually decreased before the relay releases, thus eliminating any electrical disturbances from its sudden interruption.

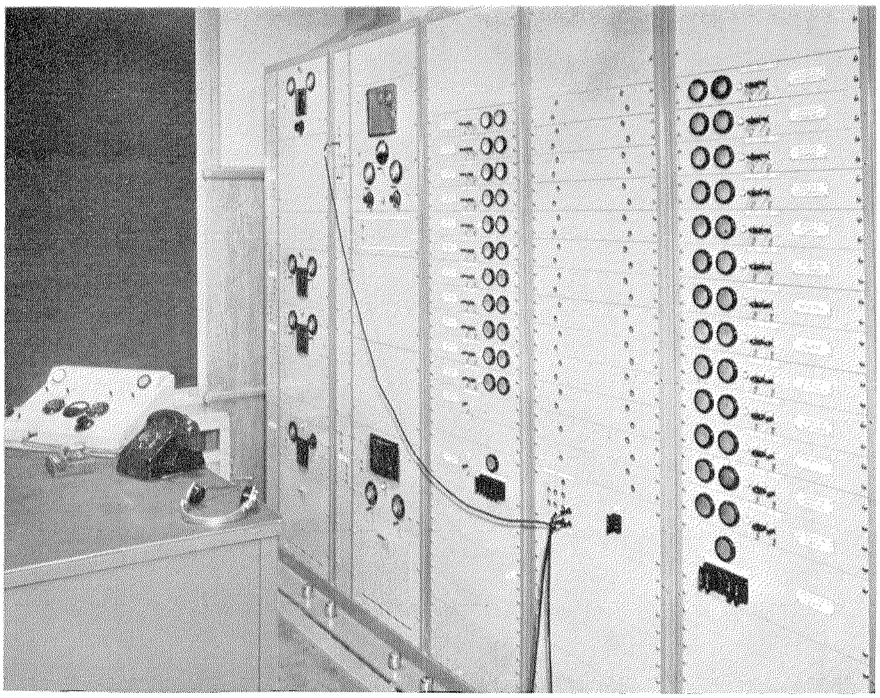


Fig. 4—Transmitter lambda. The rack at the left is for signaling apparatus and the next rack contains transmission equipment. The remaining three racks are for the alphas, each of which uses a full panel and the adjacent two jacks, which connect to the transmission channel and the program meter.

2.1 TRANSMITTER LAMBDA

To permit the broadcast supervisor to check the technical performance of a broadcast channel and radio transmitter, a high-quality fixed-tuned receiver is provided. A program meter bridged across the output of the receiver permits ready comparison with the corresponding meter from the alpha originating the program.

Fig. 4 shows the equipment racks of the transmitter lambda; the broadcast supervisor's control box is on the desk. The indicating instrument is connected to the program meter of the alpha in service. An orange lamp at the left indicates that the oscillator for transmitting the time signal is in operation and the time signal will be sent within three minutes. The time signal is transmitted automatically unless the supervisor operates a disabling switch.

2.2 ALPHA-LAMBDA FOR RECORDING

Recording has assumed a role of very great importance in modern broadcasting. The decentralized recording rooms have the same general access to lambdas for originating programs as do the alphas. Conversely, the recorder may take the place of a lambda transmitter and all alphas have access to the recorders.

As originally installed, there were five recording rooms. Two employed steel tape or "magnetophone" equipment, one recorded on cellulose-varnish discs, and another used wax masters. There was one spare room. This installation was seriously damaged during the war and is in the process of being rebuilt.

2.3 ALPHA-LAMBDA FOR MIXING

Two large mixing desks are used for complex broadcasts in which a large number of studios or other originating sources must be linked together or the outputs from which must be mixed. Their automatic switches differ from those in the lambdas in the absence of vertical locking devices. This enables the simultaneous branching of all alphas. In addition to having lines to all the alphas in the building, there are 15 junction lines for handling outside broadcasts received through the lines alpha.

On dialing, an alpha is connected immediately through its branching relay to the mixing lambda and receives confirmation of this through a green signal light. The alphas connected to the mixing lambda are automatically indicated by their numbers appearing in lights on a number board in the mixing lambda as shown in Fig. 5.

The connection from an alpha terminates at a jack on the mixing desk. The operator connects a mixing volume control to each of these jacks and gives a starting signal to the alpha desired simply by turning up the mixer control. This operation illuminates the red lamp in the alpha and changes the green signal on the number board to red.

Before operating the mixer volume control on a given alpha, the operator can listen in on that circuit by plugging his headphones into a jack and pressing a key located next to the volume control involved. This greatly facilitates the operation of the system and prevents errors.

This same system is used in mixing a program from an outside source. For convenience, the signals are also reproduced at the lines alpha from where an operator issues instructions by telephone to the remote point.

3. *Lines Alpha*

The automatic switching system is limited to transmission channels completely within the building. Programs originating outside the building must be received on manual equipment, which is concentrated in the lines alpha. For convenience, the automatic telephone switchboard for the public telephone system is also housed in this room.

To enable all possible interconnections between outside sources and the alphas and lambdas, the output circuits of all alphas and a large number of interconnecting trunks are brought into the lines alpha.

Certain auxiliary facilities are also located in this room and include equipment for receiving foreign programs, time-signal equipment, isolating amplifiers, and oscillators for checking program meters. A view of one section of lines alpha is shown in Fig. 6, the entire installation encompassing about 35 racks of equipment.

3.1 Broadcast Receiving Sets

Seven broadcast receiving sets are maintained in the lines alpha. Two of these are for high-frequency reception. A common antenna is used and consists of a single wire suspended about 15 meters (50 feet) above the roof of the building. It is connected to a broad-band amplifier, the output of which supplies signals to the seven receivers and also to the fixed-tuned receivers for monitoring the broadcast radio transmitters. The output of each receiver passes through two band-eliminating filters, which are particularly effective in suppressing certain types of interference.

The output of the receivers can be connected to the automatic program distribution system.

3.2 Time Signals

Three equipments for time signals are provided.

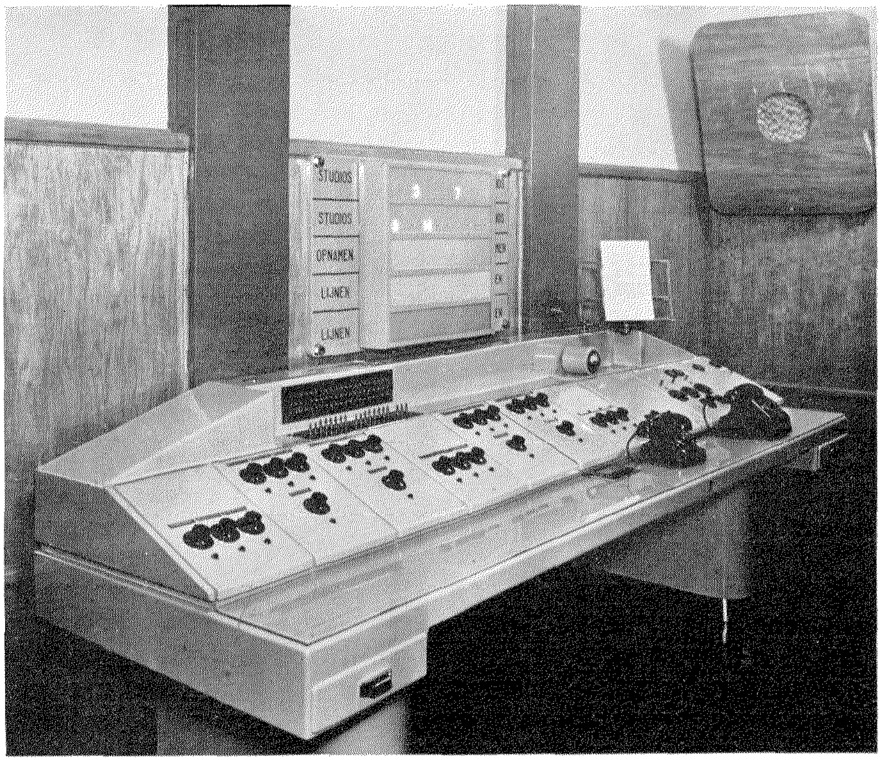


Fig. 5—Mixing-lambda desk. The number board indicates the connected alphas in green and the alpha in service in red.

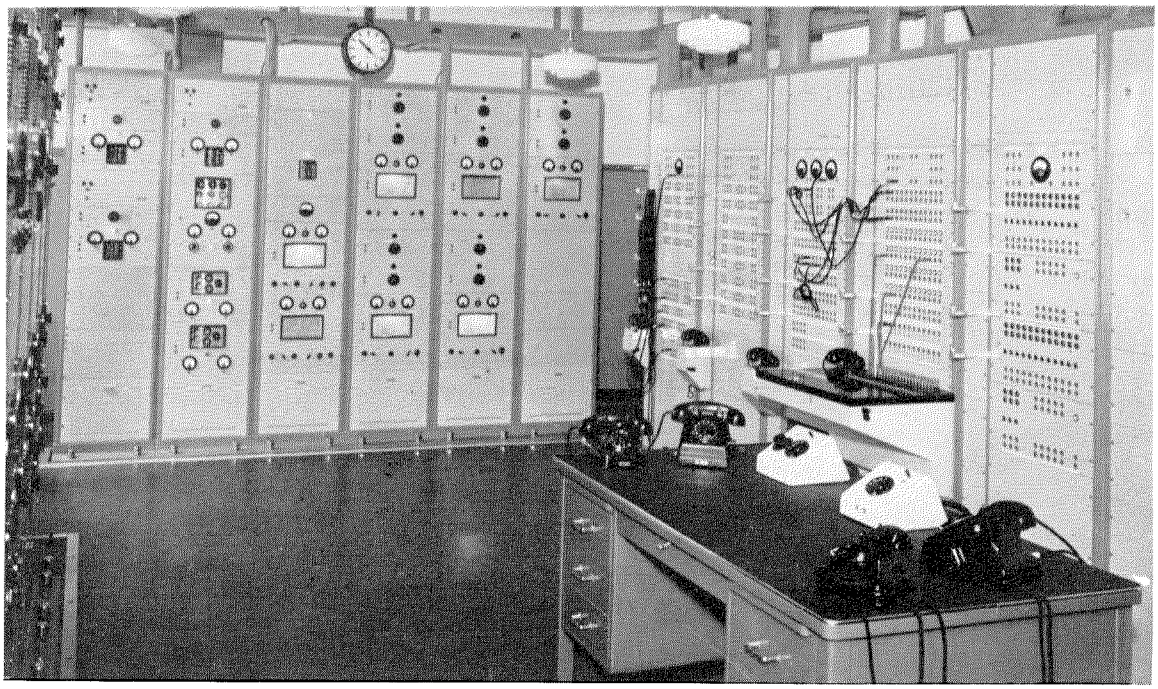


Fig. 6—Part of lines-alpha room.

These are of the "music box" type and are permanently in operation.

Being connected to the program distribution system, their output may be obtained from the listening headphone jack in each alpha. For broadcasting, the technician of the alpha connects the input of an amplifier to the proper jack and so inserts the signal into the channel.

3.3 Isolation Amplifiers

Isolation amplifiers produce no gain but are employed to prevent disturbances in circuits connected to their outputs from affecting the input circuits. About 30 of these are in constant use. For instance, all signals supplied to the program distribution system are passed through isolation amplifiers.

3.4 Program Meters

Five program meters are distributed throughout the lines alpha room to permit the program level on any transmission line to be measured. Each is provided with several indicating instruments. As each meter must be checked before it is used, two oscillators have been provided which have an output of exactly 1.55 root-mean-square volts at 1000 cycles.

4. Amplifiers, Program Meter, and Line Equalizer

4.1 AB Amplifier

The *AB* amplifier is the basic element of the speech-input installation. It consists of two push-pull resistance-capacitance-coupled stages employing triode AC2 vacuum tubes. The fine characteristics of this amplifier are a result of the care given to the individual components and the over-all design.

4.1.1 Noise

There are three principal causes of noise in the first stage of an amplifier. One of these, Johnson noise, is the result of thermal agitation of the electrons in the input resistance of the amplifier. The theoretical value of this noise voltage is equivalent for the *AB* amplifier to a noise voltage of 0.0678 inserted in the input circuit.

The electronic noise is a result of the instantaneous variations in the liberation of electrons from the cathode of the vacuum tube. This noise, converted to an equivalent voltage input to the amplifier, would be 0.0202 microvolt.

The mains hum is caused by the use of alternating current for operating the indirectly heated cathodes. For the AC2, an average hum voltage of 3 microvolts is produced for the first push-pull stage. This corresponds to a hum voltage at the input of the amplifier of 0.0428 microvolt.

The total noise voltage is equivalent at the input of the amplifier to 0.082 microvolt. As the amplifier produces less than 1-percent harmonic distortion with an input of 5 millivolts, a volume range of 95 decibels may be applied to the input.

The input transformer required particular attention from the viewpoint of minimizing noise. The transformation ratio between the line and grids directly influences the volume range of the amplifier; to reduce the noise level, the impedance ratio has been made very high, 40 to 200,000 ohms. Through careful design, the screening of this transformer at the frequencies normally giving the most trouble has been made very high.

The use of alternating current for cathode heating introduces a mains transformer. This transformer is separated from the other elements of the amplifier by a screen consisting of an aluminum sheet 2 millimeters (0.079 inch) thick sandwiched between two silicon steel sheets each 1 millimeter (0.039 inch) thick.

4.1.2 Microphonics

The problem of microphonics is essentially of a mechanical nature, the disturbances being transmitted normally either by simple mechanical shocks or by acoustic fields. The susceptibility to vibrations of the elements in the vacuum tube used in the first stage of an amplifier is the most important single factor. Although the AC2 has a very small tendency to microphonics, the wide volume range of the amplifier dictated special precautions to keep such disturbances below the noise level of the amplifier.

The effect of shocks can be eliminated by suspending the vacuum tubes individually. In the case of the *AB* amplifier, superior results were obtained when the four tubes and their coupling elements were suspended as a unit. In Fig. 7, the design of this suspension is evident. A very heavy mounting plate is supported on four long adjustable springs. Stops are provided to prevent excess motion and the entire assembly may be blocked for shipping.

Protection against acoustic fields is more difficult. It was considered necessary to put the tubes of the first stage in individual hermetic enclosures. Primarily to avoid the generation of acoustic fields within the amplifier, the transformers are impregnated with an insulating compound to hold the windings and the permalloy core laminations rigidly. It is not necessary to mount the transformers elastically.

4.1.3 Distortion

The distortion in the amplifier is almost entirely produced in the vacuum tubes of the output stage.

The push-pull output transformer is capable of operating with a 30-percent unbalance of plate currents of the output tubes with only a negligible contribution to distortion. The output transformer has an impedance ratio of 50,000 ohms to 40 ohms. Operating at a higher level than the input transformer, it requires no elaborate screening.

4.1.4 Performance

The response-frequency curve of the amplifier is flat to within ±1 decibel from 20 to 14,000 cycles. The distortion with a signal output of 1.55 volts, which is the maximum output under normal operation, is less than 0.3 percent. The distortion with an input of 5 millivolts is less than 1 percent. The phase shift is given in Fig. 8. The noise level is 94 decibels below the maximum output signal.

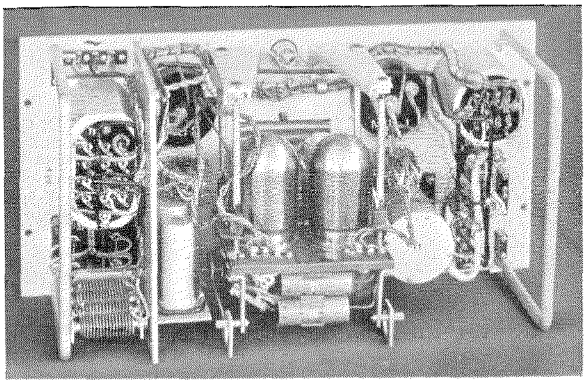


Fig. 7—*AB* amplifier showing spring suspension of tubes and coupling elements. The input tubes are in hermetically sealed cases. These precautions minimize microphonic noises.

4.2 Program Meter

The program meter fulfills the requirements set up in 1935 by the International Broadcasting Union. These requirements are that over a range from 30 to 10,000 cycles the input impedance shall be greater than 10,000 ohms and the error in reading shall not exceed 1 decibel for impulses having a minimum duration of 10 milliseconds.

The amplifier for the meter is comparable to the *AB* amplifier except that the second stage uses power triodes. The output tubes have directly heated cathodes and the noise level is 65 decibels below maximum signal. The output of the amplifier charges a capacitor through a double diode. The capacitor discharges through a dry rectifier and a resistance that is higher than that of the rectifier. The voltage across the rectifier is applied to the grid of a triode, the plate current of which is a measure of modulation depth. The resistance of the rectifier varies logarithmically with voltage giving this desirable characteristic to the program meter.

An important feature of this design is the use of uranium dioxide as the material for the resistor in series with the rectifier to provide temperature compensation. This makes unnecessary the control of temperature required in previous designs. Fig. 9 shows a front view of the program meter.

4.3 Line Equalizer

Line equalizers are required to modify the transmission characteristics of the telephone lines over which program material is transmitted to the building. These lines may be nonloaded cable or open-wire circuits. They may be loaded lines having upper cut-off frequencies varying between 2800 and 10,000 cycles. Also, they may be loaded and repeatered lines of the type used for international circuits or provided especially for broadcasting.

As all of these external circuits have a characteristic impedance of approximately 600 ohms, the equalizer has been designed for this value. A transformer converts this impedance to 40 ohms to permit the insertion of an *AB* amplifier to bring the signal to the standard value of 1.55 volts.

For the higher frequencies, a constant-impedance equalizer of the shunted-*T* type is used. The reactive circuit in the series arm consists of an inductance and capacitance in series, each of which is adjustable in 10 steps. The conjugate values of impedance provided by a parallel resonant circuit in the shunt arm are automatically selected by the switch that controls the series-arm elements.

For the lower frequencies, a comparable circuit is employed but only a simple inductive or capacitive reactance

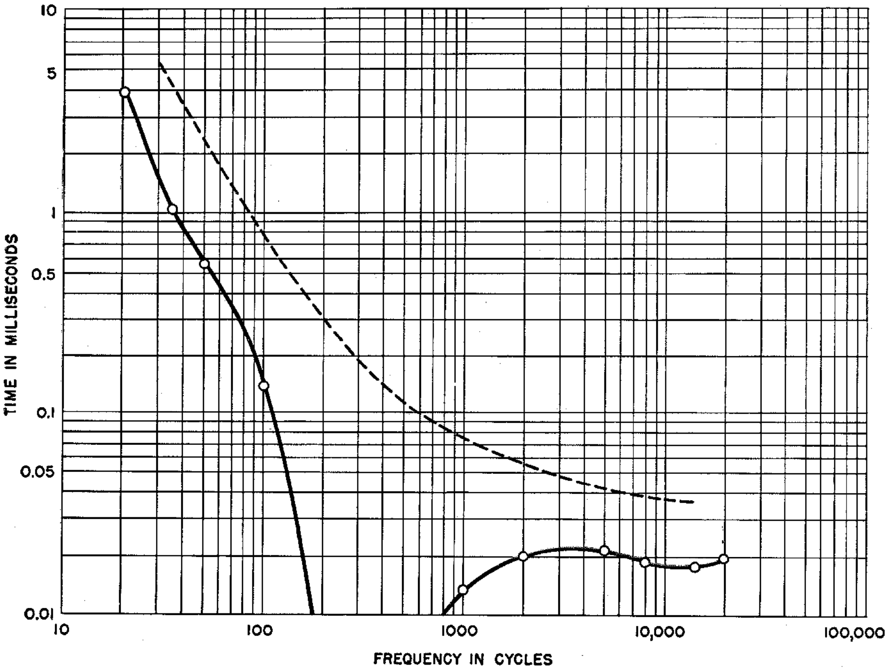


Fig. 8—Phase delay versus frequency for the *AB* amplifier for an output load of 40 ohms. The dotted curve is the maximum permissible phase delay. The solid curve is the measured values.

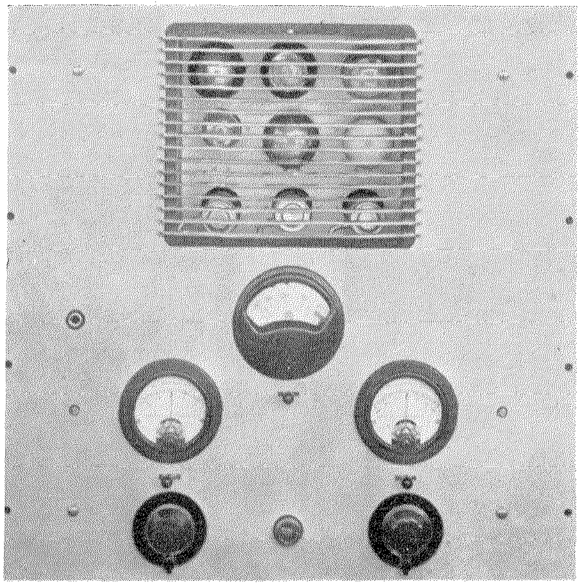


Fig. 9—Program meter.

is used instead of the resonant circuit. The reactive elements are adjustable in four steps. A switch that reverses the inductive and capacitive elements in the series and parallel arms permits either a rising or falling frequency characteristic to be obtained.

A constant-impedance attenuator having a range from 0 to 20 decibels, is provided for adjusting the input level to the *AB* amplifier.

The high- and low-frequency equalizers being independent and of constant impedance produce no interaction and the over-all equalization is the sum of the two individual equalizations. Lines may be equalized within ±1 decibel from 30 cycles to about 85 percent of the cut-off frequency of the line, and to 10,000 cycles if there is no cut-off frequency.

A line equalizer is shown in Fig. 10. The inductors use iron-dust cores. The capacitors use paper dielectric for the high-capacitance values and mica for the low values.

5. *Standardization*

In such an installation, much is to be gained by standardizing on a few basic pieces of equipment and using them wherever possible. The number of spare equipments is reduced, servicing and maintenance are simplified, and continuity of operation is improved. Consequently, considerable attention was given to these features.

5.1 CIRCUIT UNITS

The *AB* amplifier is used as a microphone amplifier, line amplifier, and in conjunction with the equalizers on incoming lines. It is also used as a reproducing amplifier in the recording rooms.

A single type of program meter is used throughout the entire installation.

A 7-watt loudspeaker-amplifier is used for monitoring programs. The amplifier is also used for recording.

Only one type of line equalizer is used.

A single type of isolating amplifier is used in the program distribution system and also for the branching of external lines.

The same scratch filter is used in reproducing records in the alphas and in the recording rooms.

5.2 VACUUM TUBES

Only three types of vacuum tubes are used extensively in the entire installation. The few other types that are employed meet special requirements that these three standardized types do not fulfill.

5.3 TRANSFORMERS

Only six types of audio-frequency transformers are used in the various amplifiers. These are the input transformer of the *AB* amplifier, the output transformer of the *AB* and of the isolating amplifiers, the output transformer of the loudspeaker amplifier, the output transformer of the program meter, the input transformer of the

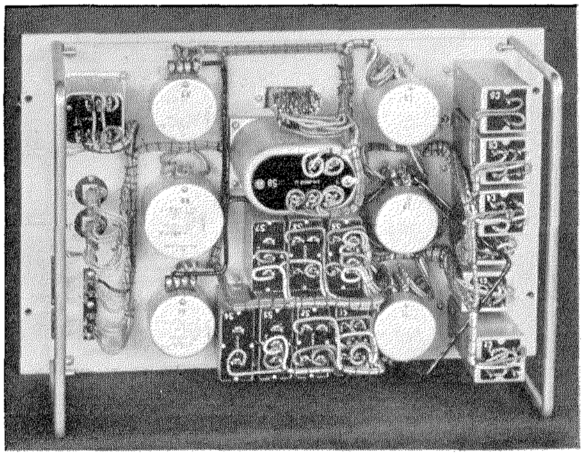


Fig. 10—Rear view of line equalizer. The control knobs are on the panel side.

isolating and loudspeaker amplifiers and of the program meter, and the input transformer of certain recording amplifiers.

In addition, only two types of mains supply transformers are used. The supply circuits of the *AB* and isolating amplifiers and of the fixed-tuned receivers are identical.

It should also be noted that all assemblies are mounted on standardized panels 480 millimeters (18.9 inches) by a multiple of 40 millimeters (1.6 inches).

6. Continuity of Service

Continuity of service is extremely important in broadcasting and much thought was given to protection against failure of equipment.

At least two microphones are available in every studio, one of which serves as a spare.

An operating amplifier in a control room can be replaced by a spare amplifier in the brief time required to shift plugs from one set of jacks to another. The spare amplifiers are normally used for secondary purposes and are thus always ready for service.

A spare volume control is provided on each control desk. If the master or main volume control becomes defective, it can be cut out of service and its function temporarily taken over by an amplifier volume control elsewhere in the channel.

Because of expense, only one program meter is installed in each alpha. In case of its failure, the alpha technician may receive instructions from the lambda to which his program is connected.

Spares are normally provided for monitor receivers, oscillators, and similar units having only single functions.

Microphone lines are always duplicated and the operator of lines alpha can readily replace a main interconnection line by one normally used for secondary functions.

7. Program Distribution System

There are a large number of individuals involved in rehearsing and broadcasting a program. Among others, they include the technical operator, producer, author, and broadcast supervisor.

There are approximately 50 sources within the building from which program material may originate. These include alphas, lambdas, recording rooms, broadcasting receivers, and talk-back and other telephone circuits. If the number of "subscribers" to the program distribution system amounts to 150, it is evident that there are 7500 possible interconnections. This is far too many to permit adequate manual switching and so automatic equipment, based on the same principles as the automatic telephone exchange, has been installed.

Each subscriber has a small desk with a calling key, signaling lamp, telephone dial, and volume control, together with a cabinet containing an amplifier and loudspeaker.

This system differs from a simple automatic telephone exchange in that most of the programs are connected to the distribution system through isolating amplifiers.

The necessity of reducing cross talk to a value 70 decibels below normal level on these circuits was a real difficulty. In a well-built telephone exchange, a cross-talk figure of more than 40 to 50 decibels is rarely attained. By using twisted and screened wires and careful arrangement of the multiple cable on the automatic switching bays, all cross-talk figures were at least 80 decibels down. In addition, all lines have been carefully balanced to earth by resistors to eliminate clicks which normally would be caused by switching operations.

8. Cabling

Broadcast transmission circuits are carried in flexible cables of single-screened pairs. Those from each alpha are bundled and terminate successively in the lines of the 10 lambdas in the lines-alpha room where they are balanced to earth by resistors.

The signaling cable of an alpha contains one quad and 15 pairs and goes directly to the automatic program distribution room. All other signaling circuits from the nontechnical rooms are directed to the nearest terminal strip from where they go by multiple-pair cables to the automatic program distribution room. The lambdas are interconnected by a signaling multiple consisting of two 15-quad cables for the green and red lamps of the alphas and two 25-

quad cables for switching relays in the transmission channels.

The automatic program distribution room is the central point of the entire cabling plant and a general distributing frame has been installed there.

For design purposes, the signaling circuits are considered as sources of disturbances and the transmission circuits as being susceptible to interference. A 25-millimeter (1-inch) twisting step was chosen for the transmission channel circuits; the twisting step for the signaling circuits, which should be a multiple of the other, was fixed at 100 millimeters (4 inches). No wire crossings or single-conductor circuits have been used.

Induced static charges have been eliminated by screening all wires and terminal strips. Screens are always earthed on the output side only to avoid setting up loop circuits.

Transmission and signaling cables are laid in separate metallic ducts, all of which are electrically linked and earthed.

In the control rooms of the alphas, microphone circuits, which are run in steel tubing, are maintained at a minimum distance of 1.5 meters (5 feet) from lighting and power circuits. The outputs of the microphone amplifiers are considered as operating at an average level and may be run in the transmission ducts without further precautions. However, they are kept as well separated as possible from cables that are operated at the reference level of 1.55 volts.

Although the entire installation is operated from alternating-current mains, the noise level is so low that the workable volume range for broadcasting is about 80 decibels. Between the large majority of the lines, cross-talk figures of over 114 decibels have been measured.

Several separate earthing systems are employed for the high-voltage power equipment, low-voltage power circuits, telephone installation, signaling circuits, and transmission circuits.

9. *Electricity Supply*

Two independent underground cables, one of which is a spare, supply power at 5250 volts to the substation of Broadcasting House. A voltage regulator maintains this supply within narrow limits. A 220-volt synchronous motor drives an alternator, which supplies power to the entire broadcasting system in the building. This arrangement provides a stabilized 50-cycle 220-volt output free from the parasitic disturbances normally associated with power delivered from a central station.

In case of failure of the city mains, the motor-generator will be driven by a Diesel engine, which starts up and couples itself automatically. During the change-over, the supply voltage to the broadcasting equipment decreases only by 10 percent for about 4 seconds. A very heavy flywheel on the shaft of the synchronous motor is responsible for this fine performance.

10. *Acknowledgment*

Started on June 8, 1935, Broadcasting House was completed on July 25, 1938. The project was directed by a Committee of Works composed of members of the Managing Council of the Institut National Belge de Radiodiffusion presided over by Mr. Marcel Malderez, at present Secretary General of the Belgian Ministry of Communications. The consulting services of Mr. Raymond Braillard, then Manager of the Control Centre of the International Broadcasting Union, were available.

I am particularly grateful to my many collaborators, among whom will be mentioned Messrs. Georges Hansen, Chief Engineer-Assistant General Manager; Gaston De Lafonteyne, Georges Gourski, and Charles Van Loo, Managing Engineers; Léo Wallenborn, now Secretary General of the International Broadcasting Organization; Eugène De Keyser and Raymond Henderickx, Senior Engineers; and Fernand Viart, Engineer.

To Bell Telephone Manufacturing Company of Antwerp was assigned the important task of constructing and installing the speech-input equipment.

Monitor for Frequency-Modulation Broadcasting*

By M. SILVER

Federal Telecommunication Laboratories, Incorporated, Nutley, New Jersey

ADEQUATE technical performance of a frequency-modulation broadcast transmitter can be obtained only through constant monitoring of the emitted wave. The center transmission frequency and percentage of modulation must be indicated at all times, and an overmodulation alarm must be provided. In addition, it must be possible to check for noise and audio-frequency distortion.

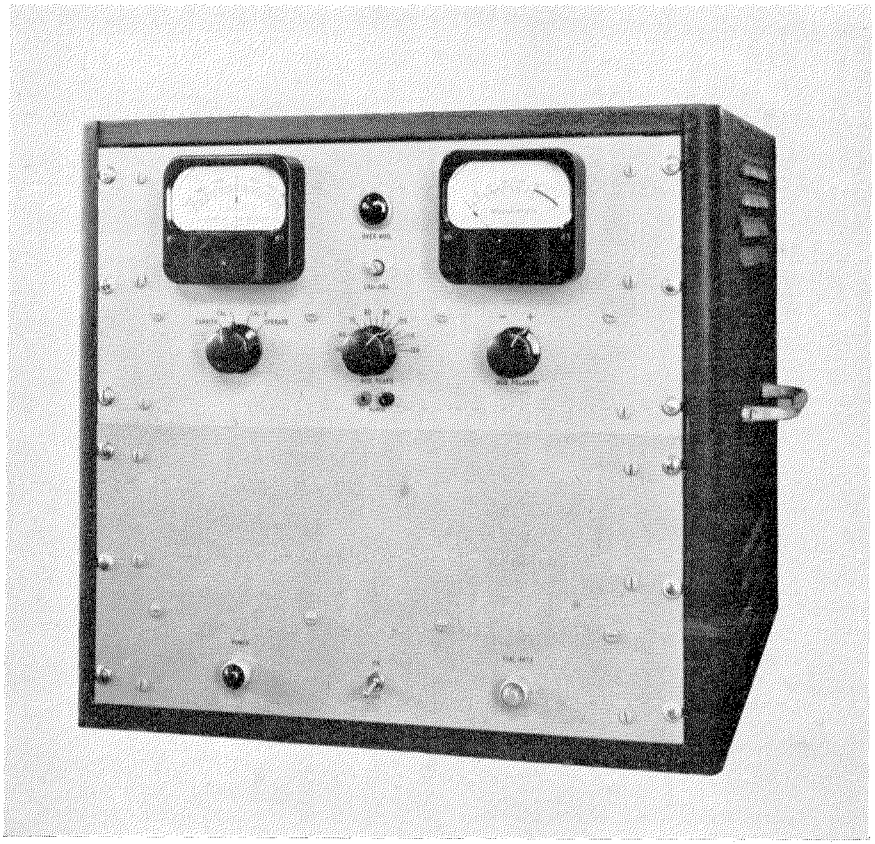
A monitor designed for this service uses a resistance-capacitance discriminator that acts as a counter to indicate the center frequency, independent of modulation deviation. Audio-frequency output from the discriminator is applied to a vacuum-tube voltmeter for modulation measurement and the operation of an alarm on overmodulation. All requirements of the Federal Communications Commission are met by this monitor.

• • •

The Federal Communications Commission requires that each frequency-modulation broadcast station in the U.S.A. have a suitable monitor for checking the frequency and other important characteristics of the transmitted wave. Frequency must be measured to within 1000 cycles per second in the band from 88 to 108 megacycles. Modulation must be measured to within ±5 percent and overmodulation must be indicated by some sort of an alarm.

Although not explicitly stated, the monitor should be capable of measuring noise and distortion of the transmitted wave. Transmitter noise must not exceed a level at least 65 decibels below 100-percent modulation and audio-frequency distortion must be less than 1 percent. The inherent noise level specified for the monitor is 75 decibels below 100-percent modulation and the inherent distortion must not exceed 0.25 percent. Another requirement is that the audio-frequency amplitude characteristic must follow a standard de-emphasis curve of 75 microseconds to within 0.5 decibel.

Several output impedance levels are required: a high-impedance output for noise and distortion measurements and 600- and 150-ohm outputs for acoustic monitoring. Both meters and overmodulation indicators should have provision for remote operation. The monitor must be self-calibrating.



Monitor for frequency-modulation transmitter operating in the 88–108-megacycle band.

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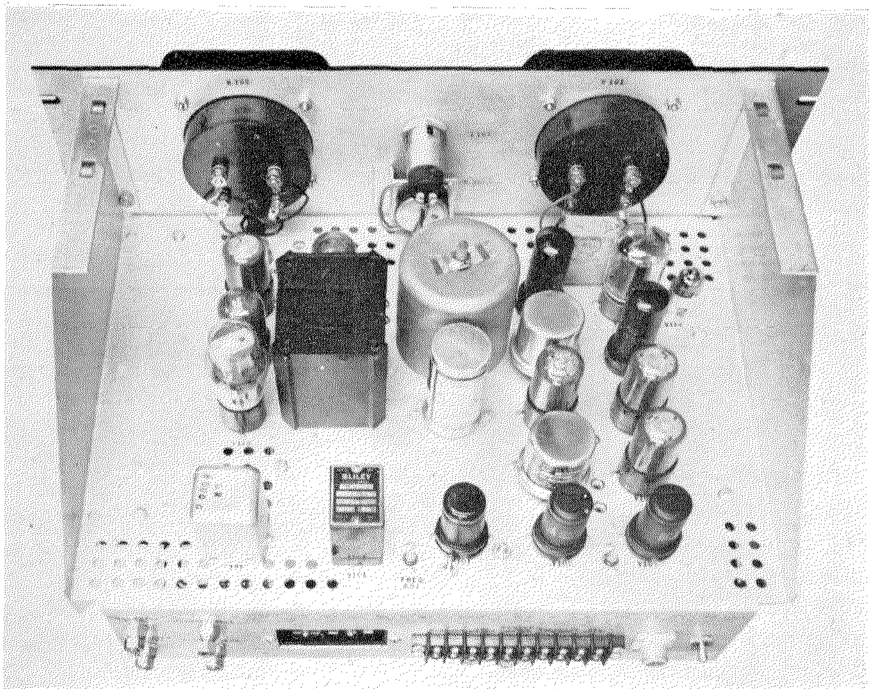
* Presented, Institute of Radio Engineers National Convention, March 5, 1947, New York, New York.

1. *Discriminator*

The discriminator is the basic element in a monitor. In general, a discriminator operating at the transmission frequency will have a greater effective band width and be better suited to modulation monitoring than if it is designed for a lower frequency. Conversely, low-frequency discriminators can be made more sensitive and stable. This suggested the desirability of using separate discriminators for these two functions.



Upper chassis of monitor. The lower chassis contains the various power supplies.

Tuned driscriminators of both single and balanced types were studied for modulation measurements directly at the transmission frequency. Satisfactory designs were developed and were capable of measuring noise and distortion of very low orders of magnitude.

A discriminator operating at 210 kilocycles was developed separately for frequency monitoring. A resistance-capacitance discriminator, it performs as a simple integrating counter. It has the necessary linearity to prevent the integration of the modulation from producing a direct-current component of sufficient amplitude to appear as an error in frequency measurement during modulation. The counter is capable of measuring frequency to within 100 cycles under conditions of 100-percent modulation.

The linearity of the counter-discriminator was far in excess of that required for modulation monitoring. The only significant problem in using it for both measurements was that of separating the 210-kilocycle intermediate frequency signal from the 15-kilocycle audio-frequency modulation. A suitable filter, having an attenuation of 60 decibels at 30 kilocycles, was devised. The final design of the monitor, therefore, uses only a single untuned discriminator.

2. *Counter*

A square wave is applied to the input grid circuit of the counter tube shown in Fig. 1. This square wave is derived from a series of cascaded clippers and limiters. When the negative square wave is applied to the grid, the counter-tube plate current is completely cut off. The capacitor *C*1 is charged during this period through rectifier 1 to the peak value of the plate voltage. During the positive half cycle the tube is conductive and the capacitor discharges through rectifier 2

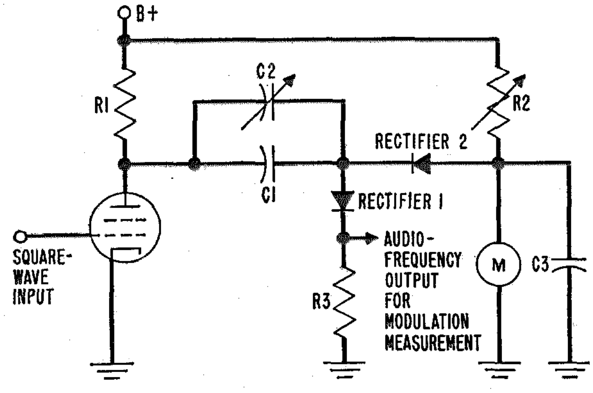


Fig. 1—Counter-discriminator circuit.

and the counter tube. The average value of current flowing through the meter circuit is proportional to the rate of discharge of the capacitor and is, therefore, a measure of frequency. In addition, the voltage across resistor $R3$ is a measure of frequency and is proportional to the modulation. This permits the same discriminator to be used for monitoring frequency and modulation.

Two adjustments are provided on the counter, one for the quiescent balance of the meter indicating frequency and the other for the sensitivity of the counter. The variable resistor $R2$ controls a bucking voltage applied across the meter M. This adjustment is used for the initial balance of the counter. The charge capacitor is made variable, $C2$, to permit adjustment of sensitivity, the amount of current flowing through the meter M being proportional to capacitance.

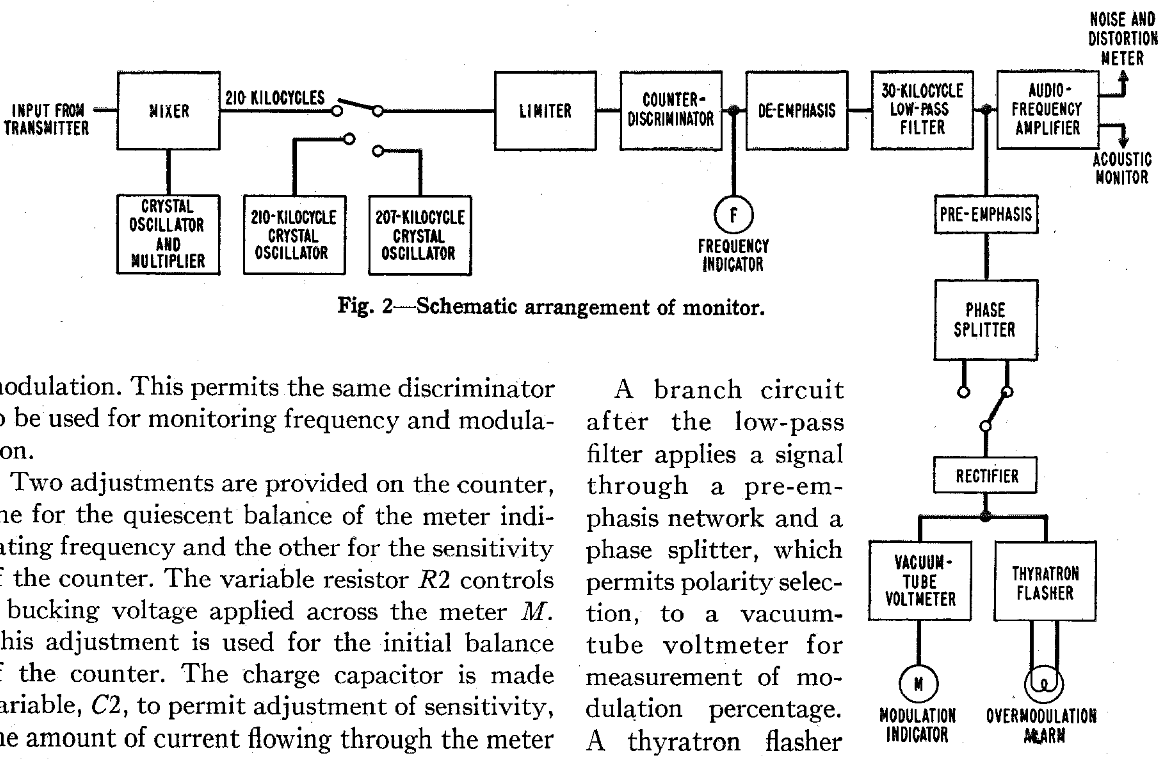


Fig. 2—Schematic arrangement of monitor.

3. Monitor

Basically, the monitor consists of cascaded limiters, counter-discriminator, filter network, calibrating circuits, and modulation-indicating devices as indicated in Fig. 2. In operation, energy from the transmitted wave is mixed with a crystal-controlled oscillation to produce a difference frequency of 210 kilocycles. The crystal oscillator is temperature stabilized and is the operating standard for frequency measurements.

The 210-kilocycle output of the mixer is applied to the counter-discriminator tube through a series of cascaded limiter circuits, which square and limit the voltage. Frequency is indicated by the meter in the counter circuit.

The output across $R3$ of Fig. 1 is applied through a standard de-emphasis network to a cathode-follower operating into a low-pass filter having a cut-off frequency of 30 kilocycles. The signal then passes through an amplifier having 600- and 150-ohm balanced outputs for acoustic monitoring and a high-impedance termination for a noise and distortion analyzer.

A branch circuit after the low-pass filter applies a signal through a pre-emphasis network and a phase splitter, which permits polarity selection, to a vacuum-tube voltmeter for measurement of modulation percentage. A thyratron flasher circuit indicates overmodulation.

Two crystal oscillators are provided for calibrating the discriminator. One supplies a 210-kilocycle signal to permit zero adjustment of the counter-discriminator for frequency measurement. The second produces a 207-kilocycle signal for adjusting the sensitivity (discriminator slope) of the counter circuit.

3.1 LIMITER

The limiter circuit is conventional in all respects but one. If all circuits are of the plate-limiting type, the limiter is a high-gain amplifier and subject to oscillation. To avoid this possibility, a clipper circuit was used in the first stage followed by a conventional plate-limiting second stage.

3.2 Modulation Indicator

Fig. 3 is a schematic diagram of the modulation-indicator circuit. The audio-frequency signal from the counter-discriminator is rectified and applied to the grid of a cathode-follower, which serves as a branch of a bridge circuit. A quiesent balance is obtained by adjusting *R*3. Under modulation, the balance is disturbed and a current, which is proportional to the amplitude of modulation, flows through the meter circuit. The sensitivity of indication may be varied by *R*2.

The same rectifier is used to actuate the thyratron flasher circuits. Alternating voltage is applied to the plate of the thyratron so that plate current is interrupted on the negative alternations. The characteristics of the rectifier circuit are such that the charge time is much shorter than the discharge time and peak indications persist for an appreciable period after a modulation peak has passed. When the voltage on the grid of the thyratron exceeds a gate value determined by the adjustments of *R*4 and *R*5, the thyratron tube becomes conductive causing a lamp in the plate circuit to glow. *R*1 is a sensitivity control.

The voltage to the filaments of the bridge and thyratron tubes is regulated by a ballast tube. This stabilizes the modulation indicator under line-voltage fluctuations. In addition, the screen circuit of the thyratron is arranged to compensate for changes in plate supply voltage as a function of line voltage.

To calibrate the modulation-indicating circuits, a 60-cycle signal is applied to the input of the rectifier circuit. The amplitude of this alternating voltage is set at the factory to correspond to 100-percent modulation. Provision is made for checking the amplitude of this voltage in operation.

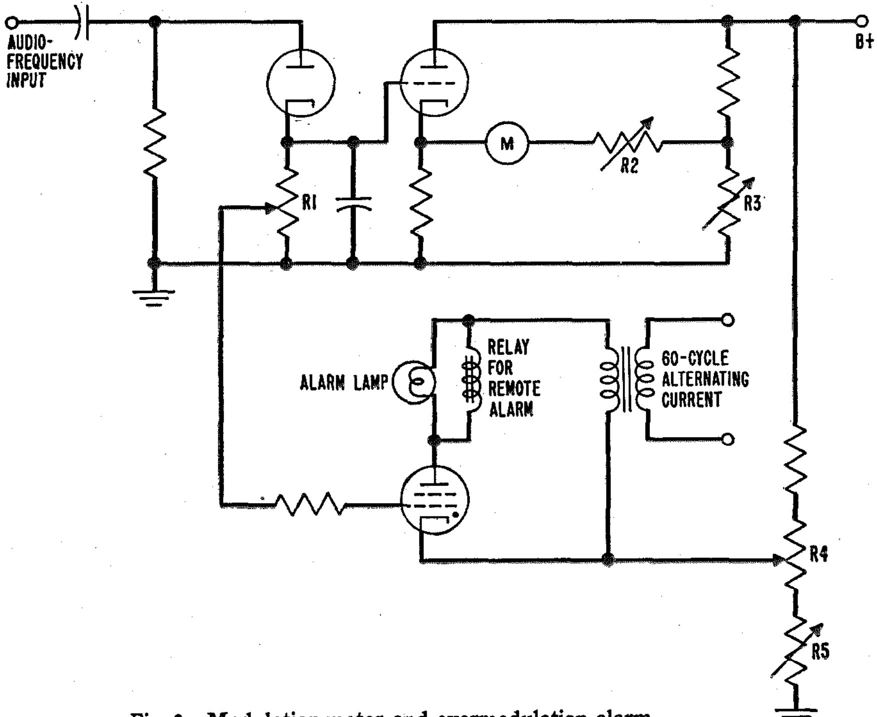


Fig. 3—Modulation meter and overmodulation alarm.

4. *Measurement of Monitor Characteristics*

Measuring the characteristics of a monitor is somewhat of a problem, and it was thought better to use indirect means of measurement rather than build a super monitor for the purpose. Suitable standards are available to measure the accuracy of frequency calibration, so this feature will not be discussed.

The most difficult characteristics to determine are inherent noise and distortion. Consider the measurement of inherent noise. With the output of the monitor connected to a suitable noise and distortion analyzer, there was applied to the monitor input a signal that was frequency modulated at a 400-cycle rate with a deviation of ±75 kilocycles. A broadcast transmitter was used as a signal source. This establishes a reference level for 100-percent modulation on the noise and distortion analyzer. If there is substituted for the signal source, a crystal oscillator using direct current for plate and filament supply, the residual noise of the monitor may be obtained. Fortunately, such a circuit exists within the monitor in the form of the 210-kilocycle calibration oscillator. The noise level measured in this manner is 80 decibels below that of the reference level, and is better than that specified.

Determining the inherent distortion is somewhat more complex. The distortion inherent in the counter circuit and in the audio-frequency amplifier circuits must be known. The counter distortion was computed by means of a static plot of frequency versus output direct voltage of the counter. With an accurate potentiometer, voltages were measured to four places. The distortion in the counter was found to be in the order of 0.05 percent. The distortion inherent in the audio-frequency amplifier was measured by the use of a noise and distortion analyzer using an audio-frequency oscillator whose distortion is 0.1 percent. A distortion of 0.2 percent was measured at 50 cycles, the frequency at which distortion was maximum. The distortion of the discriminator being so much lower than that of the audio-frequency amplifier, the inherent distortion of the monitor was determined as 0.2 percent, which is below the permitted value.

Summing up the characteristics of the monitor, the inherent noise is 80 decibels below a reference of 100-percent modulation. The inherent distortion is 0.2 percent. The error in frequency indication under conditions of 100-percent modulation is ±100 cycles, assuming proper calibration of the reference crystal oscillator. The error in modulation percentage and overmodulation indication is ±5 percent.

It is interesting to note that a plate power supply with a ripple level of less than 2 millivolts was required. A hard-tube regulator circuit with additional feedback was used. It was also necessary to use direct current for the filament circuits to reduce the noise level to the low values measured. The filament supply uses selenium rectifiers. The complete equipment, including all power supplies, is contained in two chassis.

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Medium-Power Multichannel Communication Transmitters

By B. T. ELLIS

Federal Telephone and Radio Corporation, Clifton, New Jersey

POINT-TO-POINT and ground-to-aircraft communication normally requires two or more channels in different parts of the frequency spectrum for uninterrupted service throughout the day and year. To avoid uneconomical duplication of equipment, a transmitter has been developed that has one power supply and modulator but two radio-frequency units. Thus, a radiotelephone and radiotelegraph or two radiotelegraph circuits can be operated simultaneously at two different radio frequencies. Four additional radio-frequency units may be added to permit any two of six channels to be so utilized. Output powers of 400, 500, and 200 watts are produced in the bands from 200 to 540 kilocycles, 2 to 20 megacycles, and 108 to 140 megacycles, respectively.

• • •

It is well known that in the high-frequency spectrum the optimum operating frequency for a given point-to-point circuit varies widely with such changing factors as season of the year, time of day, local atmospheric conditions, and the degrees of utilization of the available channels. Transmitting equipment, therefore, should be capable of convenient frequency change or separate transmitters may be maintained for each of the operating frequencies. A very effective solution to this problem is the use of a multichannel transmitter in which several radio-frequency units, each permanently adjusted to a single frequency, are operated from a common power supply. Flexibility and speed of frequency change are thus combined with economy and efficiency.

The *FTR*-3 was the first transmitter built on the multichannel principle. Radio-frequency units, power supplies, and modulators are constructed in separate cabinets, which can be set up side-by-side in a sectionalized frame to meet individual requirements. Radio-frequency units and modulators can be controlled remotely, the desired facilities being selected by operating a telephone dial. This philosophy has been widely followed by other manufacturers.

The *FTR*-3 is rated at an output of 3 kilowatts from each radio-frequency unit. A similar transmitter, the *FTR*-5, is rated at 5 kilowatts per channel. Transmitters 184, 185, and 186, which are described here, provide output powers in the fractional-kilowatt range.

1. General Design

The new transmitters depart from the horizontal side-by-side arrangement of their predecessors. As lower power permits much smaller basic components, the units may be positioned vertically as well as horizontally, with a saving in floor space. Cabinets have been designed to house several combinations and to provide the necessary ventilation, metering, and interconnecting circuits.

Transmitter output is nominally rated at 400 watts for the 200- to 540-kilocycle unit, 500 watts for the 2- to 20-megacycle unit, and at 200 watts for the 108- to 140-megacycle unit. All units are capable of 100-percent amplitude modulation. Primary power is 220 volts, 50 or 60 cycles, single phase.

All major units, except high-voltage rectifiers, are constructed on movable chassis, which may be slid part way out of the cabinet for inspection. Flexible cables maintain electrical connections to units thus withdrawn. Supply voltages and control circuits are brought to each unit by a cabinet cable assembly that includes a large terminal board for connection of external control equipment. Links and jumpers, adjusted by the user, connect each radio-frequency unit for telephone or continuous-wave operation, the choice of which may be made either directly or remotely.

2. Cabinets

The basic cabinet assembly is the left-hand half of the transmitter shown in the illustration. The high-voltage rectifier is on the floor of the cabinet, the filter components are on a sliding shelf over the rectifier, and a modulator chassis slides into place over the filter. Two radio-frequency units fit side-by-side in the top section. Space is provided between the modulator and radio-frequency units for control equipment and low-voltage and bias rectifiers.

The basic assembly permits simultaneous operation on two channels, either with two telegraph or with one telegraph and one telephone circuit. Only one modulator is provided, thus limiting telephony to a single channel at a time. Two identical units will double this service. Another arrangement is shown in the illustration. It uses a single power supply and modulator with up to six radio-frequency units, any two of which may be operated simultaneously.

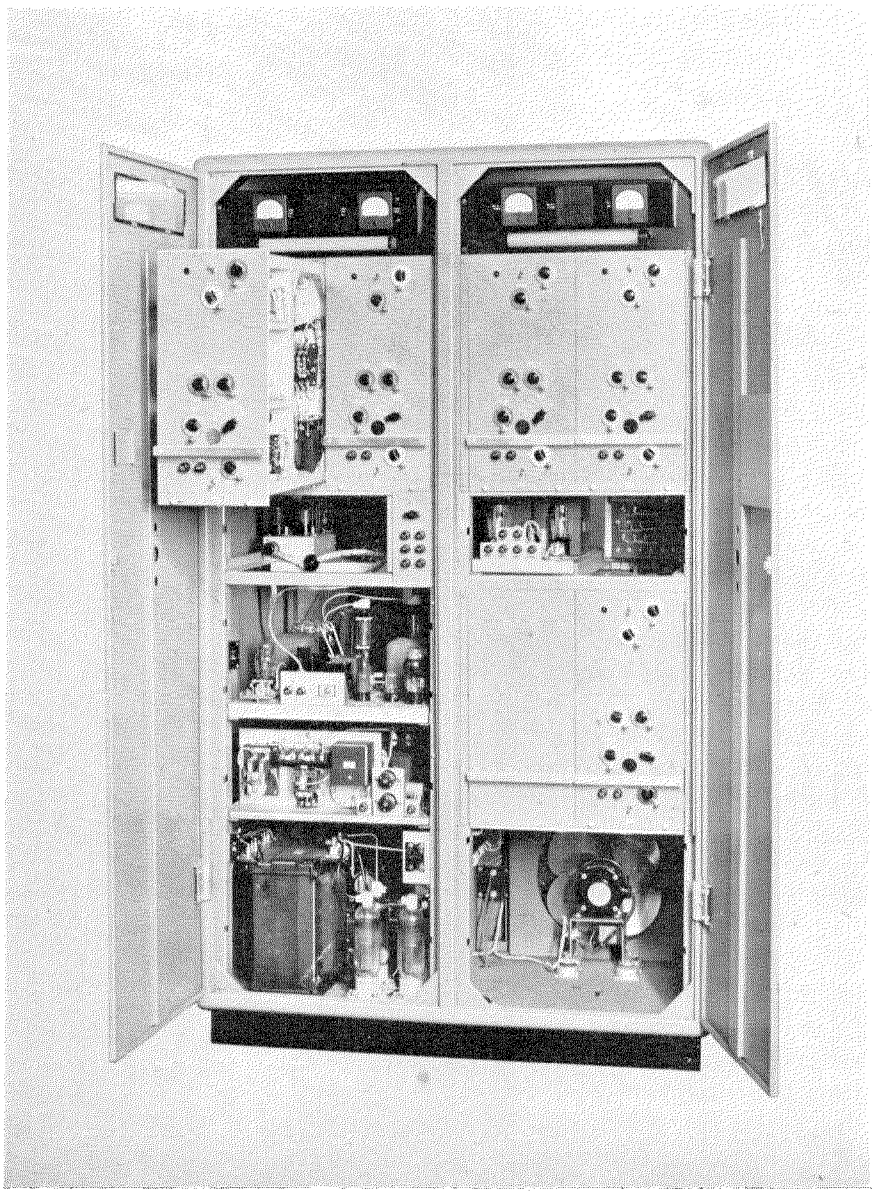
3. *Power Supply*

Two radio-frequency units, one of which may be voice modulated, can be operated simultaneously from one power supply. When more than two radio-frequency units are associated with a single power supply, only two may be on the air at the same time.

The main high-voltage supply delivers 1500 volts. This voltage was selected as a compromise between high plate efficiency and size of components. The rectifier is a center-tapped bridge circuit using F-872A tubes. 750 volts is available for the intermediate stages. Tuned-choke filters, adjustable for 50- and 60-cycle power frequencies, are used on both the 1500-volt and the 750-volt outputs. Bias voltage (−300 volts) and low-voltage plate power (+300 volts) are obtained from two rectifiers each using 5U4G tubes and single-section filters.

4. *Radio-Frequency Units*

Three basic radio-frequency units cover the frequency ranges 200 to 540 kilocycles, 2 to 20



Two basic units are mounted in a single cabinet. The left-hand assembly contains a power supply at the bottom with its filter just above; then come the modulator, bias and low-voltage power supplies, and two radio-frequency units. Four additional radio-frequency units may be accommodated in the right-hand assembly. The subassembly panels have been removed to show the general arrangement of components. In operation, the full doors are closed, the meters being visible through the windows at the top.

megacycles, and 108 to 140 megacycles. They are interchangeable in the cabinet. The low- and very-high-frequency units have 72-ohm coaxial output transmission lines while the high-frequency unit is provided with a 600-ohm spaced parallel-wire line. The tube complements for the three types of radio-frequency units are listed in Table I.

TABLE I

VACUUM-TUBE COMPLEMENTS FOR RADIO-FREQUENCY UNITS

Function	200–540 Kilocycles	2–20 Megacycles	108–140 Megacycles
Oscillator	6AC7	6AC7	807
Buffer	6V6	—	—
1st Multiplier	—	6V6	807
2nd Multiplier	—	2E22	815
Intermediate Power Amplifier	2E22	2E22	815
Power Amplifier, Two Tubes	152TH	152TH	4–125A

Crystal-controlled oscillators are used in all radio-frequency units, although the circuits may be modified easily to permit frequency-shift keying or variable-frequency oscillator operation.

Blocked-grid keying is used in the low- and high-frequency units to permit keying speeds up to 500 words per minute. The very-high-frequency unit is not intended for radiotelegraph operation.

Frequency range adjustments are made with movable links and flexible leads that are soldered to coil taps. Original designs provided tap switches to permit bandsetting from the front panel, but most applications require so few frequency changes during the life of the equipment that the initial cost and the maintenance expense of band switches are not justified.

The low- and high-frequency units use a novel construction of intermediate stages. Each tube is mounted upside-down in an inverted U-shaped bracket so as to project through the shelf. By-pass capacitors, grid leak, and parasite suppressors are mounted on the top and sides of the bracket with very short leads to tube socket terminals and with considerably improved accessibility for test and replacement. The shelf and bracket is produced as a subassembly to facilitate manufacture.

5. *Modulator*

Approximately 350 watts of audio-frequency power is available for modulation of one radio-frequency unit. The input level for 100-percent modulation is 25 decibels below 6 milliwatts in the 500-ohm input circuit. The frequency response is designed for voice transmission. Low-pass input and output filters may be inserted to reduce higher-order sidebands. Distortion, without the low-pass filters, is less than 10 percent at 100-percent modulation.

All the audio-frequency amplifier stages are push-pull triodes. Resistance coupling is used between the 6SL7GT first amplifier, 6SN7GT second amplifier, and 6B4G driver tubes. The modulator stage uses two 805 tubes operating class *B*. The use of triodes in the driver stage makes inverse feedback unnecessary.

6. *Acknowledgment*

Development work on the equipment was directed by J. W. Butterworth who is now Chief Engineer of the Marine Division of Mackay Radio and Telegraph Company.

Modernization of International Telephone Service, and Its Reaction on National Telephone Systems*

By E. P. G. WRIGHT

Standard Telecommunication Laboratories, Limited, London, England

THE present state of the international telephone service is of interest because pre-war techniques are, in many cases, obsolete and because, in any event, dislocations caused by the war necessitate some revision.

This paper gives general information about foreign telephone systems, outlines continental networks and their interconnection, and describes future planning procedure and the progress already made.

Specialists in international telephony believe that the traffic will expand quickly with a rapid and efficient service. To provide higher speed, manual operation must be replaced by semi- or fully automatic methods. Consequently, existing national services need reviewing and, if necessary, replanning as integral parts of the international system. The modernization of some foreign national systems is discussed, with particular regard to quicker service.

There is striking evidence that harmony between nations is likely to be obtained only by mutual understanding, and that international communication should form one of the principal ties upon which the future of civilization will depend. It is probably unavoidable that national and international projects will be needing expenditure simultaneously, and it will be a tragedy if the expansion of the international service is stifled in order to allow a disproportionate effort on the national services.

• • •

1. Existing European International Service

1.1 DESCRIPTION OF THE SERVICE IN 1939

During the 20 years between the two world wars, periodical reports were made of the introduction of new international circuits, permitting the public an ever-increasing range over which telephone calls could be established.

The service which has grown up has many admirable features, but a searching examination shows that progress has been hampered by the circumstances existing in the years during which the service was growing.

It would be expected that the international telephone service would benefit from its ability to adopt all the better practices that were in use by different administrations and to avoid any failures that were found to exist in national systems. In actual fact, these benefits have not accrued, and there has been a tendency for the international service to be less efficient than the best service provided nationally.

This result is perhaps not so unexpected when it is realized that the development, trial, acceptance, and introduction of a new practice require the agreement of two or more groups of technicians for an international service, whilst only one group is involved for a national service. Secondly, there is rather more scope for the introduction of improved methods in a national service, where replaced equipment can be retired for use on subsidiary circuits having less exacting requirements, but this convenience is seldom so easily planned in connection with international circuits. Thirdly, the difference in language and custom in different countries sometimes rules out the simplest solution and results in the continuance of a well-tried, but inefficient, practice. Finally, there is the important fact that an efficient international service cannot be established on the basis of several different practices, and many administrations would prefer to operate all their different international circuits on a uniform practice, even though this were not particularly efficient, rather than mix several methods, some good and some indifferent.

1.2 FUNCTION OF THE C.C.I.F.

In spite of the handicaps already mentioned, the interests of the international service are fostered by those who, by their foresight and wisdom, have created on a sturdy, yet resilient, basis the Comité Consultatif International Téléphonique, (C.C.I.F.), whose object is the coordination of the international service, and by those who have helped to ensure the successful working of this association, believing that its objectives would benefit the different peoples of the world. It is the function of the Comité Consultatif International Téléphonique to think in terms of international service, disregarding purely national matters. In general terms, this work includes the investigation and study of new problems and specific instances of difficulty with existing practice, the establishment of standards for engineering design and procedure, operating and maintenance practice, documentation, statistics, and nomenclature.

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* Presented, Institution of Electrical Engineers, London, England, March 20, 1947, and Norwegian Society of Electrical Engineers, Oslo, Norway, September 29, 1947. This paper received the Fahie Premium of the Institution of Electrical Engineers.

All the large administrations in Europe are members of the Comité Consultatif International Téléphonique, and most of the larger administrations in North and South America are represented, in addition to several in other parts of the world.

This association, which has been established for more than 20 years, is not dissimilar to the United Nations Organisation in its wideness of outlook. It maintains a permanent Secretary-General, secretarial staff, and reference laboratory, and all expenses are met by fees paid by the co-operating members in proportion to their size. It is abundantly clear that the Comité Consultatif International Téléphonique provides the machinery for maintaining the international service in a healthy and vigorous condition.

1.3 Causes of Service Delays

Not unnaturally, the international service has grown stage by stage as technical knowledge and the prospect of a reasonable financial return have enabled more and more international circuits to be built. It is an undisputed fact that this manner of growing does not lead to the best result, as is demonstrated by the delay experienced by the public in obtaining international calls. Most of the principal cities of the world have automatic telephones, and local calls can usually be established in about 20 seconds; international calls, on the other hand, often require 1, 2, or even 4 hours, and there is no insoluble technical reason for such a serious disparity. The delays can be attributed to an inadequate provision of circuits, to the operating procedure, and to the primitive nature of signalling facilities. To a lesser extent the service may be restricted by the variable quality and instability of the circuits.

An examination of these causes shows clearly that replanning is essential. For example, in 1938, out of a possible 406 pairs of countries in Europe, 121 pairs were interconnected, with approximately 4 circuits per route; there were 316 subsidiary groups, with an average of less than 2 circuits per route; in three instances there were as many as 30 separate groups between adjacent countries, with less than 2 circuits per group.[1] In such circumstances, inefficiency is unavoidable—either the subscribers are made to wait so that the circuits are reasonably loaded or, if delay is to be eliminated, the circuits will not earn a satisfactory revenue. With large quantities of small groups, a satisfactory service is almost beyond the control of the administration. In most cases, the traffic increases and the delay grows simultaneously, until a point is reached when the delay is so great that it stifles further increase of traffic. It is evident that the continuance of these small groups does not easily admit the possibility of an adequate revenue without a hampering service delay.

The most obvious rearrangement to operate a smaller number of larger groups involves the introduction of additional switching points, and it has been judged in the past that the creation of switching points is a greater evil than the use of small groups. The limitations of manually operated switching centres are well known, and detailed operation procedures and carefully trained operators are necessary to obtain even a normal grade of service. The handling time of a call with manual transit-switching is several times that of a direct call; if two such switching points are involved, a further increase in time is introduced. Bitter experience has resulted in a widespread disinclination to adopt manually switched calls whenever a direct service is possible.

This state of affairs may be largely explained by the following facts:

A. The designation of the international call is longer than that of the national call and therefore more difficult to memorize.

B. The charge to the subscriber must be based on a ticket including all relevant information.

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[1] "Nomenclature des Circuits Téléphoniques Internationaux" (Bureau de l'Union Internationale des Télécommunications, Berne, February, 1939).

C. The passing of the call order from operator to operator and the further recording of the information is slow on account of the length of the designation and because at some point language translation may be necessary.

D. The prospect of finding three circuits in series—all available simultaneously—is poor, and when finally this is achieved there is an increased difficulty in finding the calling party because it is likely that he has moved from one place to another while waiting for the call to mature.

These difficulties are accentuated by the fact that it is the invariable practice to use ringers for signalling and that the facilities thus provided are equivalent to those on magneto switchboards, which have long been obsolescent on all the major national circuits.

2. Possibility of Improving Service

2.1 APPLICATION OF RECENT DEVELOPMENTS TO INTERNATIONAL SERVICE

The use of automatic switching and modern line construction and signalling methods in the national networks has shown that a quick and accurate service can be provided, and superficially there is no reason why a corresponding application of new methods to the international networks should not be equally effective.[2]

2.2 WHEN TO PLAN MODERNIZATION

Although such a proposal as that in Section 2.1 may be the logical conclusion, there remains the question whether this is the correct time to plan a general revision of the international service. Once a new plan is approved, it will take several years to introduce, and it is therefore desirable that it should remain in operation for at least 25 to 30 years.

During this time, technical progress will be made, and it is clearly inadvisable to exclude the benefits of any major developments which may become available. On this account, the system planner needs to scrutinize the situation and consider the possibility of:

A. New and cheaper line construction by means of wave-guides, coaxial cables, or radio.

B. The introduction of new switching means, such as the use of electronic methods.

C. Modifications to the subscriber's instrument to introduce higher-quality or auxiliary services, and the possibility in either case of consequential reaction on the frequency band transmitted.

D. Rearrangement of the basis of subscriber numbering or dialling.

E. Rearrangement of the basis of tariff or method of recording charges.

F. Introduction of associated services.

The outcome of such an examination can scarcely be precise, and, although few technicians would agree that research on these subjects will be unproductive during the next 30 years, there is little indication that anything is to be gained by delay from the research point of view. Recent development work with coaxial transmission systems has been carried to the point where it has been established that an efficient service is available for large groups of long lines at a reduced cost. Further new systems do not appear to offer the possibility of such striking improvements which would be universally applicable within the next decade.

The amount of switching equipment that will be needed at automatic transit centres is relatively small, and for this reason special arrangements can be made to provide desirable facilities, which could not be attempted if the scale were more extensive. Furthermore, the relatively small amount of equipment involved for switching means that it would not be a serious problem to replace it completely if new developments of sufficient importance became available. Any new international plan will be influenced by modern ideas on numbering, dialling, and charge recording.

The argument in favour of replanning in Europe is strengthened by the devastation caused by the war. Many of the international circuits passed through Germany, Berlin being the principal switching centre for eastern Europe. The rebuilding of the European international system is unavoidable, and there is everything to gain by replanning it on a world-wide basis.

2.3 NEED FOR INTERNATIONAL ENTERPRISE

Assuming that the technical merit of a new international switching plan could be proved to satisfy the critics, and that it could be established that the volume of traffic was likely to increase in a predictable manner with improved service, there still remains a doubt whether there exists sufficient international enterprise to permit the logical expansion to take place. This aspect cannot be overlooked, because, although there is a potential demand for a better international service on the basis of existing charges, the new plant needs a capital expenditure requiring Government financial approval. In the process of seeking such approval, this international enterprise will be in competition with a number of national projects backed by the national Press or other eager supporters. If it is appreciated that the fundamental need for this enterprise lies in promoting international understanding—and, as such, it is a moral obligation comparable with the accepted principles of the United Nations Organisation—there is a reasonable chance of progress; but, if the case is judged on a narrower issue, who is there in a position to expostulate with the financial authority who sees but small advantage in promoting the project? If it is a fact that a general plan of co-operation between nations has a better chance of obtaining financial approval than a series of individual plans, this provides a considerable argument in favour of such a co-operative international plan, because of its greater prospect of surmounting adverse nationalistic or political interests.

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[2] H. T. Kohlhaas, "Milestones of Communication Progress," *Electrical Communication*, v. 20, n. 3, pp. 143–185; 1942.

3. *Telephone Service Throughout the World*

3.1 Survey of International and Intercontinental Services

In order to gain perspective, it is wise to study the question, What proportion of the international service is provided in Europe? The answer seems to be larger than might be expected.

The largest number of telephone calls originate in the United States of America, which has a network of international circuits to other parts of the world. A great volume of traffic passes between the United States and Canada, and probably more international calls pass over this frontier than over any other. The network of circuits to Mexico, the West Indies, and Central and South American states is probably roughly equal to the network of radio circuits to different capitals in Europe. Both Argentina and Brazil have direct radio connection to a number of capitals in Europe. Argentina has a network of land lines and radio circuits to other states in South America, but the quantities are all relatively small. Table I gives an approximate summary of intercontinental circuits in 1939.

TABLE I

Intercontinental Circuits in 1939

	North America	Oceania	Central America and West Indies	South America	Asia	Africa
Europe	25	1	—	14	3	5
North America		9	6	9	—	—
Oceania			—	—	2	—

Note: These figures are based on reconnections recommended by the 13th Plenary Assembly of the Comité Consultatif International Téléphonique (London), 1945.

In Europe at present, there are some 28 independent states, and in 1939 there were some 600 different international routes, many of which passed through intermediate territories. The service between Great Britain and France was provided by a group of more than 100 circuits. It is probable that the negotiations necessary to operate the European traffic between so many administrations, involving as it must the complications of many languages and many currency systems, needs more effort than corresponding work in America, where a very large proportion has been negotiated between two administrations on the basis of one language and one currency.

It is evident that the technical problems in America are very great on account of the long distances involved, but to a large extent these difficulties have been overcome by the development work of the American Telephone and Telegraph Company and the Bell Telephone Laboratories to meet national requirements. Although an association like the Comité Consultatif International Téléphonique has immense technical resources, it is easier in some ways for a single organization, working independently, to develop and introduce a new service.

As regards international networks in other parts of the world, it is noteworthy that some of the long-distance national circuits between states in the Commonwealth of Australia (as shown in Table II) are of equivalent length to other international circuits, but, although the technical problems are equivalent to those encountered in

Europe, the technical administration for handling problems and deciding on policy is centralized as in the U.S.A.

South Africa and the Middle and Far East contain important telephone centres, but their international service constitutes a very small percentage of the whole. For example, India has a community of interest with Burma to the east and Ceylon to the south. The Soviet Union covers an immense territory, partly in the European network, but it seems unlikely that the eastern portions will have needs extending beyond adjacent countries, which will be determined directly by the administrations concerned. This applies also to China; and the international traffic from Japan is not extensive. The establishment of many of the standards for international service is likely to be settled independently in Europe and North America for land lines and switching, while for radio circuits initial standards are likely to be agreed mutually between these two continents.

TABLE II

Australian Inter-State Toll Cable Routes*

Circuit	Equipments × channels per equipment	Distance in kilometres
Adelaide–Darwin	1×3	3 050
Adelaide–Melbourne	1×12, 8×3, *and* 5×1	770
Adelaide–Perth	10×3 and 8×1	2 625
Brisbane–Sydney	2×12, 20×3, and 7×1	1 060
Melbourne–Sydney	4×12, 18×3, and 5×1	933

* Longest and largest routes in use or undergoing installation.

3.2 Outline of Development Proposals in North America

Proposals for the ultimate development of a continental switching plan for North America have been published. In general, it is hoped to treat the United States and Canada as a whole. As a means for improving the long-distance service, facilities will be provided for each long-distance operator to have access, at her own control, to every subscriber connected to a machine-switching office. This plan envisages the construction of a national numbering plan, whereby any subscriber may be called from any part of the country by using the appropriate long-distance number. In a sense, this is true at the present time, but the essential significance of the national numbering plan is that each number can be called by the operator on her dial or keyset. This procedure means that the name of the town and state must be replaced by a number or code letter, and obviously each telephone network must have a characteristic code. The plan provides not only for operator connection but also for subscriber dialling in the future; this does not necessarily anticipate the complete elimination of long-distance operators, but it is certain that subscribers will be given facilities for dialling over an increasing area, and it is very desirable that a national numbering plan should be applicable to both operator and subscriber dialling.

One of the difficulties involved in subscriber dialling is to find a long-distance dialling code that will identify the class of call as rapidly as possible and route the call independently of the local network. It is undesirable to arrange local registers to deal with long-distance traffic and nearly as wasteful to switch a call initially to a local register and then to a long-distance register.

In America, it is proposed to choose long-distance codes including the digits 1 and 0, which never appear in existing multi-office codes. Three-digit codes, of which the second digit is 1 or 0, will provide for some 75 zone centres which, it is thought, will suffice for the United States and Canada. It is intended that this code should be retransmitted from point to point until the terminal zone centre is reached, and by this means the choice of alternative routes will be greatly simplified. The crossbar system provides an excellent opportunity for rapid switching at the transit centres, and the large toll switching office at Philadelphia gives a practical example of the speed of service obtained with keyset sending, call announcing, and high-speed switching.[3–6]

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[3] F. J. Scudder, and J. N. Reynolds, "Crossbar Dial Telephone Switching System," *Bell System Technical Journal*, v. 18, p. 76; 1939.

[4] B. C. Bellowes, "Philadelphia Adopts Automatic Toll Switching," *Bell Laboratories Record*, v. 22, p. 101; 1943.

[5] L. G. Abraham, A. J. Busch, and F. F. Shipley, "Crossbar Toll Switching System," *Electrical Engineering*, v. 63, p. 302; 1944.

[6] F. F. Shipley, "Nation-wide Dialling," *Bell Laboratories Record*, v. 23, p. 368; 1945.

4. *Perspective of European International Service as Seen from Great Britain and the Continent*

The geographical position of Great Britain is unique, and the absence of a land frontier has not seriously obstructed modern methods of communication, because the largest group of international circuits lies between London and Paris, a distance of about 450 kilometres. This is the shortest international route from London, but it is much longer than the shortest route from many other countries. Fig. 1 illustrates this fact; it represents an outline of Great Britain laid over the north-west of Europe, with the south coast of England lying along the coasts of Belgium and The Netherlands. The important continental towns and countries have been shown in their relative positions within this outline in order to emphasize their proximity to one another. It will be observed that the area of England and Scotland contains portions of France, Belgium,

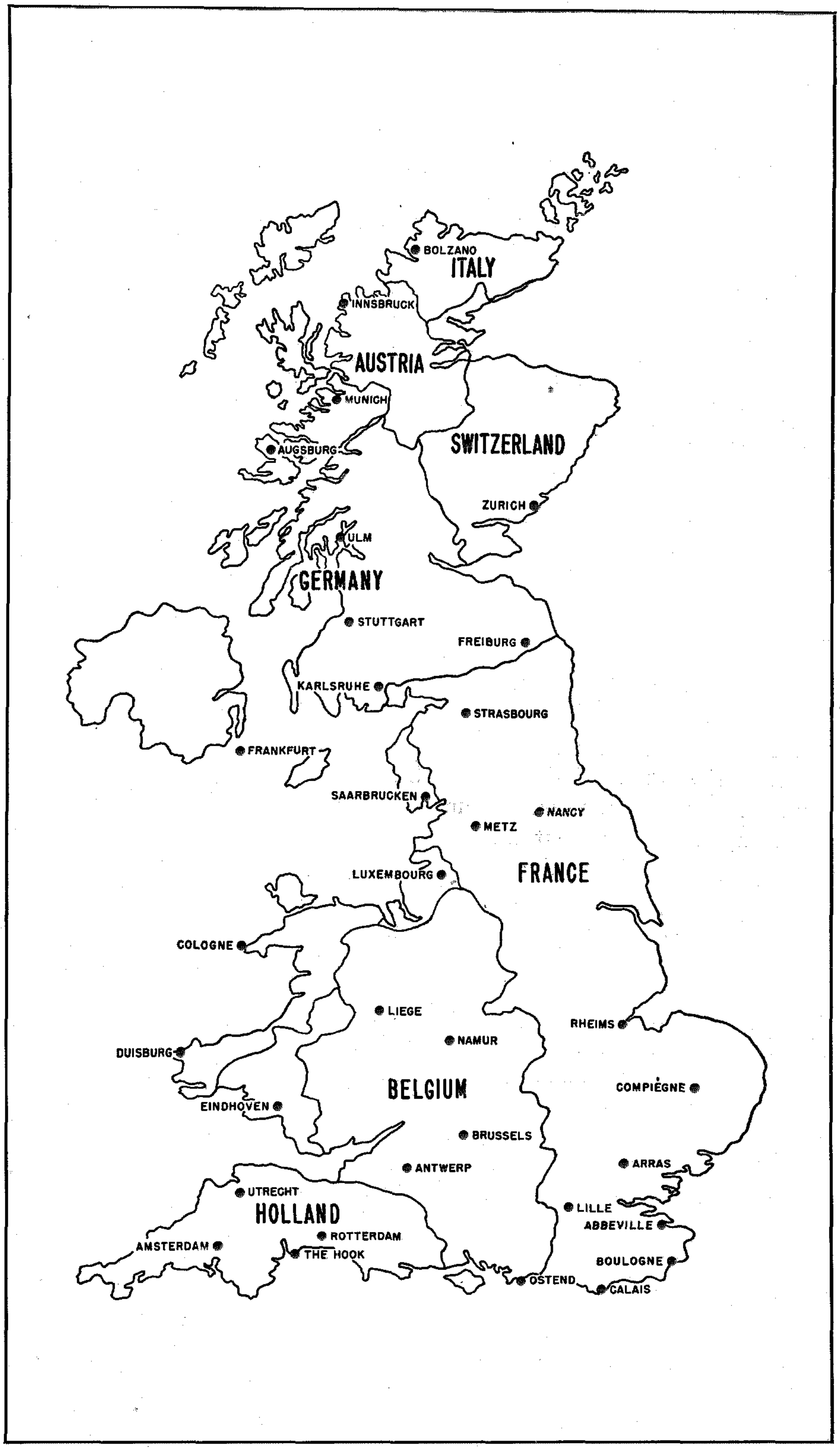


Fig. 1—Superimposition of map of British Isles on the continent of Europe, illustrating close association of European countries.

Holland, Luxembourg, Germany, Switzerland, Italy, and Austria.

The proximity of all the towns in this area to their nearest frontier is striking, and in many cases international calls involve a shorter distance than a trunk call in Great Britain. It must be realized that, consequently, the need for subscriber dialling on international calls is likely to arise sooner on the continent than in this country.[7]

5. *Revised European Fundamental Switching Plan*

5.1 NATIONAL NUMBERING PLAN

The fundamental switching plan in Europe is likely to follow principles similar to those outlined for North America, but elementary differences are likely to arise. In America, the continental switching plan is being designed to reach the wanted subscriber, whilst in Europe the international switching plan will be designed to take the call to the international *tête-de-ligne* centre of the country required, the completion of the call from this point being a national matter. In America, a single numbering plan will embrace the whole of the United States and Canada, but in Europe each country will wish to set up an independent national numbering plan. The designation of the called number must be transmitted over the international circuit, and it is of importance to know whether the number of digits is variable and, if letters are used as well as digits, the number of impulses to which the letters correspond.[8]

Table III shows the number-letter combinations in New York, London, Paris, Berlin, Sydney, and Copenhagen. The relationship between the dial numbers and the corresponding impulses also needs consideration. Fig. 2 compares the dial in most general use with that used in Sweden.

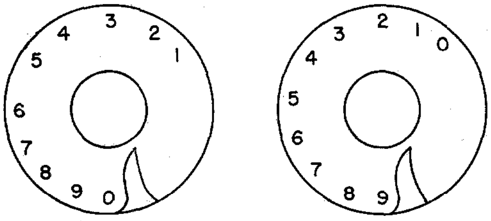
The letter-number variation is more difficult to allow for than the variation in the number ring for which compensation can be arranged in the national incoming register.

It seems very desirable that all the administrations in Europe should have national numbering plans. When this is achieved an international call can be passed into a country from any direction and then extended to the wanted subscriber over the national network without the intervention of an incoming operator. This fact is best illustrated by the case of Germany, where traffic from the surrounding countries would be presented to the international *tête-de-ligne* centres at Hamburg, Cologne, Frankfurt, Stuttgart, Münich, Berlin, and Stralsund. With a national numbering plan, each of these centres would have access to all the others for the national service, and the international traffic could be merged with the national traffic at these centres.[9] It might be thought that for countries having only a single international *tête-de-ligne* centre—such as Great Britain, Norway, or Portugal—the national numbering plan would be less essential, but such is not the case, because, if the service is to be based on the number being set up by the originating *tête-de-ligne* operator, it is necessary that she should be able to call this number on her dial or keyset, and this require-

TABLE III

NUMBER-LETTER COMBINATIONS

Numbers	Letters					
	New York	London	Paris	Berlin	Sydney	Copenhagen
1	—	—	—	A	A	Central
2	ABC	ABC	ABC	B	B	ABD
3	DEF	DEF	DEF	C	E	EFG
4	GHI	GHI	GHI	D	J	HIK
5	JKL	JKL	JKL	E	L	LMN
6	MNO	MN	MN	F	M	OPR
7	PRS	PRS	PRS	G	U	STU
8	TUV	TUV	TUV	H	W	WXY
9	WXY	WXY	WXY	J	X	AEØ
0	Z	O	OQ	K	Y	—



GENERAL SWEDISH

Fig. 2—Comparison of dial number plates.

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[7] "Long Distance Telephony," *The Post Office Green Papers*, no. 43; 1938.

[8] "Comité Consultatif International Téléphonique—Report of Sub-Commissions' Meeting in Paris, June and July, 1946," *Post Office Electrical Engineers' Journal*, v. 39, p. 117; 1946.

[9] E. Muller-Mees, "Automatic Methods of Trunk Operation used by the Reichspost," *Europaischer Fernsprech Dienst*, v. 60, p. 29; 1942.

ment necessitates a unique combination of digits for each subscriber.

It is possible to use a national numbering plan for setting up calls before full dialling facilities are actually available, because it enables the originating operator to set up the call without passing the designation verbally to an inward operator. No very formidable difficulties are expected in the layout of national numbering plans. Switzerland, Belgium, The Netherlands, and Denmark have national numbering plans wholly or partly in use at present; France, Italy, Rumania, and Hungary have plans prepared; and Sweden has a numbering plan covering the more populated part of the country. (See Section 8, Appendix.)

It is apparent that the use of national numbering plans will entail some further education of subscribers. More frequent access to directories is not necessary, but subscribers must be encouraged to include both their national and local telephone numbers in their correspondence letter-heads; usually the national number will be the local number with the addition of a 2- or 3-digit prefix. This prefix will serve to replace the name of the town, and subscribers must learn to ask for the country and the national number, and not the country, town, and local number. This seems to be a complication, but in Switzerland and Belgium it has not caused any serious trouble. It is obvious, of course, that the same national number should be applicable to both the international and the national services.

5.2 GENERAL TECHNICAL REQUIREMENTS

Although the introduction of national numbering plans is possibly not essential for a semi-automatic international service, it is certainly very desirable, and it seems likely that most administrations will adjust their national service to enable the international service to benefit. There are, however, other subjects which need to be studied in connection with a semi-automatic international service, e.g., consideration of international traffic, fundamental cable plan, fundamental transmission plan, speed of connection, signalling system, service tones, language differences, tariff, alternative routes, and numbering plans. Each of these subjects deserves extensive examination.

5.2.1 Consideration of International Traffic

Attention has already been directed to the fact that a large number of small groups results in delay, and one of the chief benefits that should emerge from a new cable plan is a reduced number of groups, each containing a larger number of channels. The cable plan must be in harmony with the international traffic originated by each country. This traffic can conveniently be divided up into three parts, namely, traffic between neighbouring towns across a national frontier; traffic to adjacent countries; and traffic which must pass through intermediate countries.

The first part will probably be handled by frontier circuits limited to terminal traffic only. There will be cases where two towns relatively close together cannot be satisfied by frontier circuits limited to terminal traffic, and in such cases direct or transit international circuits will be needed. Copenhagen and Malmö are good examples. The second part contains about 60 per cent of the whole; the cables for this service can be fairly readily planned, as in most cases they will need to join the capital towns and the route will be indicated by existing cables. These routes will have a bearing on the third part, which can be treated in several ways, and the choice of the most suitable arrangement depends on many separate issues needing examination.

If the cable plan is to depend chiefly on the transit traffic, the anticipated requirements of this type of traffic should be set down as follows:

A. Switching should be rapid—if possible as rapid as that of the direct circuits.

B. Provision of alternative and emergency routes.

C. The operator should not need to know whether the service is direct or indirect.

D. Large traffic groups are preferable to small.

E. Merging with national traffic is undesirable on the transit circuits.

F. The use of the cable routes provided for the direct routes is advantageous.

5.2.2 Fundamental Cable Plan

Apart from the provision of frontier circuits and direct circuits as justified, there are three possible ways of preparing the cable plan:

A. The provision of one central transit centre with cables radiating in all directions to every country. The

TABLE IV

Forecast of European International Traffic for 1952

	Austria	Belgium	Bulgaria	Czechoslovakia	Denmark	Finland	France	Germany	Great Britain	Greece	Hungary	Italy	Luxembourg	Netherlands	Norway	Poland	Rumania	Spain and Portugal	Sweden	Switzerland	Turkey	U.S.S.R.	Yugoslavia
Austria	—																						
Belgium	3.8	—																					
Bulgaria	0.25	0.05	—																				
Czechoslovakia	(*a*) 22 (*b*) 30	2	0.5	—																			
Denmark	0.5	2	0.1	2	—																		
Finland	0.3	0.05	0.05	0.25	2	—																	
France	7.5	(*a*) 60 (*b*) 30	0.25	6	2	0.5	—																
Germany	20	20	0.25	20	10	1	(*a*) 15 (*b*) 15	—															
Great Britain	4	30	0.25	2	8	0.5	80	21	—														
Greece	0.25	0.05	4	0.25	0.1	0.05	0.2	0.25	0.25	—													
Hungary	40	2	0.33	16	0.25	0.25	3	8	4	0.33	—												
Italy	(*a*) 12 (*b*) 10	2	0.25	3	0.25	0.1	(*a*) 10 (*b*) 12	20	14	0.5	7	—											
Luxembourg	0.2	12	0.05	0.5	0.1	0.05	(*a*) 5 (*b*) 5	4	0.1	0.05	0.05	0.05	—										
Netherlands	4	(*a*) 25 (*b*) 25	0.05	6	6	0.1	36	(*a*) 32 (*b*) 30	29	0.05	3	4	2	—									
Norway	0.25	0.1	0.03	0.5	5	1	1	4	6	0.03	0.1	0.15	0.05	0.75	—								
Poland	6.4	2	0.1	6	4	1	4	(*a*) 15 (*b*) 7	4	0.03	4	4	0.15	2	2	—							
Rumania	5	0.1	2	10	0.1	0.05	4	3	2	1	(*a*) 7 (*b*) 7	1.5	0.05	0.05	0.05	2	—						
Spain and Portugal	0.5	0.1	0.05	0.1	0.1	0.05	(*a*) 5 (*b*) 5	2	0.2	0.05	0.1	0.2	0.1	0.1	0.05	0.1	0.05	—					
Sweden	0.5	2	0.05	2	24	15	4	16	0.5	0.05	1	0.2	0.2	4	(*a*) 15 (*b*) 8	2	0.1	0.1	—				
Switzerland	20	10	0.1	4	2	0.1	(*a*) 38 (*b*) 30	(*a*) 35 (*b*) 20	28	0.05	8	(*a*) 40 (*b*) 40	2	26	1	4	0.1	4	2	—			
Turkey	0.25	0.05	3	0.25	0.1	0.05	0.2	0.25	0.25	4	0.33	0.25	0.05	0.05	0.03	0.05	1	0.05	0.05	0.05	—		
U.S.S.R.	3	0.25	0.5	2	0.25	1	2	5	2	0.25	2	1	0.05	0.5	0.1	(*a*) 3 (*b*) 3	2	0.25	1.5	0.2	0.15	—	
Yugoslavia	11	0.1	4	3	0.1	0.05	2	3	1	5	12	8	0.05	0.1	0.03	2	5	0.05	0.05	4	1.5	2	—

All traffic expressed in erlangs (call-hours). The traffic in both directions is assumed to be equal. Letters (*a*) and (*b*) indicate main and frontier traffic, respectively.

TABLE V

Summary of Cable Plan Studies for 23 Countries in Europe, Showing Comparison Between Star and Ring Formations

	Austria	Belgium	Bulgaria	Czecho-slovakia	Denmark	Finland	France	Germany
Star Formation								
Direct circuits	156 350	102 140	20 700	132 930	70 540	22 650	366 440	(323 265)
Overflow from transit centre	9 450	1 870	1 560	10 300	4 540	700	11 670	(See
Indirect circuits	6 430	13 210	6 640	11 620	10 880	6 420	8 680	Note
Circuits to transit centre	8 000	19 460	18 700	7 920	9 540	18 480	22 320	2)
	180 230	136 680	47 600	162 770	95 500	48 250	409 110	323 265
Ring Formation								
Direct circuits	65 670	32 620	17 800	54 440	19 140	13 590	89 310	270 820
Terminations from ring	75 200	12 080	3 760	39 400	9 590	2 090	52 360	39 130
Ring circuits	33 040	87 720	1 350	58 270	54 820	5 120	238 520	680
Circuits to ring	—	—	14 250	—	—	24 000	—	—
	173 910	132 420	37 160	152 110	83 550	44 800	380 190	310 630

	Gt. Britain	Greece	Hungary	Italy	Luxem-bourg	Nether-lands	Norway	Poland
Star Formation								
Direct circuits	326 730	32 880	132 030	181 450	9 330	190 710	60 710	139 760
Overflow from transit centre	5 740	1 569	8 700	11 100	2 920	3 150	4 840	7 960
Indirect circuits	7 890	6 640	8 690	13 190	3 690	6 550	3 210	2 100
Circuits to transit centre	15 600	27 500	24 400	30 000	11 000	9 880	21 600	20 400
	355 960	68 580	173 820	235 740	26 940	210 290	90 360	170 220
Ring Formation								
Direct circuits	12 160	17 800	44 880	78 760	7 180	42 090	50 440	50 690
Terminations from ring	43 210	3 270	20 600	22 700	1 510	38 640	2 930	17 430
Ring circuits	120 080	1 350	52 170	39 600	3 470	100 180	12 330	94 720
Circuits to ring	136 250	26 920	35 270	39 960	10 150	—	21 210	—
	311 700	49 340	152 920	181 020	22 310	180 910	86 910	162 840

	Rumania	Spain and Portugal	Sweden	Switzer-land	Turkey	U.S.S.R.	Yugoslavia	TOTALS
Star Formation								
Direct circuits	77 430	37 200	110 720	306 300	25 500	80 380	69 700	2 768 570
Overflow from transit centre	8 440	1 830	3 510	3 380	1 560	6 690	11 150	122 620
Indirect circuits	9 610	5 040	6 280	5 660	6 640	10 930	8 390	375 665
Circuits to transit centre	33 250	24 000	14 000	10 800	26 950	50 400	31 200	455 400
	128 730	68 070	134 510	326 140	60 650	148 400	120 440	3 722 255
Ring Formation								
Direct circuits	33 220	6 000	50 690	59 800	17 800	35 170	50 700	1 120 770
Terminations from ring	6 780	4 040	14 250	81 360	3 270	9 670	4 870	508 140
Ring circuits	17 070	6 200	60 040	165 200	1 350	26 390	5 490	1 185 160
Circuits to ring	50 450	39 950	—	—	20 400	45 500	43 000	507 310
	107 520	56 190	124 980	306 360	42 820	116 730	104 060	3 321 380

Note 1: All quantities represent circuit-kilometres.

Note 2: The star formation total for Germany includes 207 275 circuit-kilometres radiating from Berlin and 115 990 circuit-kilometres from other centres in Germany. The former group combines with the "overflow" and "indirect" groups from other countries and is included in the "indirect" total.

Note 3: The circuit-kilometre totals should not be given too great a significance. The annual charges will depend very largely on the number of circuits per route.

Note 4: The route-kilometres involved are as follows:

Star Formation		*Ring Formation*	
Direct	260 000	Direct	90 000
Overflow from transit centre Indirect through transit centre	27 000	Termination from ring	26 000
		Ring circuits 2 × 4 600 =	9 200
		To ring	26 000
To transit centre	27 000		

Note 5: Dividing the circuit-kilometre totals by the lengths given in Note 4, the following loading rates are obtained:

Star Formation		*Ring Formation*	
Direct	$\frac{2\,769\,000}{260\,000} = 10.65$	Direct	$\frac{1\,121\,000}{90\,000} = 12.45$
Overflow and Indirect	$\frac{498\,000}{27\,000} = 18.4$	Terminations	$\frac{508\,000}{26\,000} = 19.5$
To transit	$\frac{455\,000}{27\,000} = 16.85$	Ring	$\frac{1\,185\,000}{9\,200} = 128.8$
		To ring	$\frac{507\,000}{26\,000} = 19.5$

cable plan in 1939 approximated to this arrangement, Berlin being the transit centre for eastern European traffic.

B. The provision of a number of primary transit centres, each having direct circuits to all other transit centres, and secondary transit centres having lines to at least one of the primary transit centres. Such an arrangement limits to two the international transit centres in any connection. The limit of one primary-to-primary circuit and two primary-to-secondary circuits provides a ready means of regulating the maximum permissible transmission loss between the originating and terminating *tête-de-ligne* centres. This type of plan seems likely to be adopted for the United States and Canada.

C. The provision of a transit ring arranged to connect a number of regional centres. Regional or national centres not lying on the ring would be given access to at least two of the regional centres on it. The length of the ring circumference and its position would be indicated both by the international telephone traffic and by the best natural cross-country routes. These factors have led to the establishment of the following routes in Europe: Paris–Brussels–Rotterdam–Copenhagen–Malmö; London–Rotterdam–Copenhagen–Malmö; London–Paris–Zürich–Milan; Paris–Zürich–Vienna; and Vienna–Brno–Cracow–Warsaw.

The international routes for traffic from Oslo, Amsterdam, Rotterdam, London, Brussels, Paris, Zürich, Madrid, Milan, and Rome to Berlin may not be reconstructed, because the traffic to Germany can be terminated at closer zone centres.

The choice of site for each transit centre should be influenced by the desirability of nominating not more than one town in each country (preferably one, such as Malmö, close to the frontier nearest to the centre of Europe).

For purposes of comparison between the single-transit-centre plan (Case A) and the transit-ring (Case C), a forecast of the international traffic for 1952, based on the somewhat abnormal traffic of 1938, is given in Table IV. The comparison is based on a central transit centre at Berlin and a ring connecting Malmö, Copenhagen, Rotterdam, Brussels, Paris, Zürich, Vienna, Brno, Cracow, and Danzig.

Table V gives a summary of the circuit-kilometre requirements of a number of countries. It is noteworthy that the direct circuits are often of use to only two countries, whereas terminations to or from the transit centres are common to a greater extent and the circuits around the transit ring are of general use. It is also to be noted that the transit ring shows a considerable circuit-kilometre saving for countries outside the ring. The total length of each class of circuit is also included, because this enables the size of the different groups to be calculated. The length of the transit ring is shown doubled in order to allow for two traffic rings, one used in each direction.

The transit ring appears to have several technical advantages, namely, economy of initial cost, less switching at each transit centre leading to faster switching, and facilities for extensions to be made more readily. From the point of view of stability of service, the ring provides a means of setting up emergency circuits, and with two routes to the ring there should be much less chance of interruption of service. The provision of a ring calls for the co-operation of a number of administrations, and it may be easier for many of the administrations to obtain funds to build a portion of an international transit ring than to obtain the same amount in order to reduce delay on existing international routes.

It is apparent that the single transit centre suffers because outlying national centres are too remote, and it follows that some saving can be introduced by adding more transit centres. Suppose, for example, that the capitals of the largest countries are chosen, i.e., London, Berlin, Paris, and Moscow. Paris can be useful as a distribution centre for the extreme west, but neither London nor Moscow can help to reach the north, the south or the south-east. A more promising rearrangement would be the choice of the most suitable geographical centres for these regions, e.g., a site in Sleswig for Scandinavia, a site in Switzerland to serve the south and south-west, a site in Czechoslovakia or eastern Austria to serve the surrounding region, and a site on the lower Danube basin to reach the Balkans. Such a modification transforms the single-transit-centre plan into the primary-transit-centre plan. However, the latter differs from the transit ring only by virtue of the fact that, with the ring, full interconnection between all the regional centres is not specified. In Europe, there is a tendency for the international telephone traffic to be inversely proportional to distance, so that it is found that in several cases over 95 per cent of the outgoing traffic does not need to pass over more than two ring transit centres. This condition is demonstrated in Table VI, which shows the volume of traffic from each of the named transit centres to each of the others. It is apparent that there is no case for interconnecting all the transit centres by direct circuits, and in fact it seems that in most cases direct circuits to two other centres (in each direction) are sufficient. The ring gives access to a number of zone centres in Germany, and it is assumed that any call to Germany may enter by any of these zone centres.

If the international traffic in Europe were more widely dispersed, there would be a greater appeal in the primary-transit-centre plan, but there would still remain the problem of whether a regional centre could be agreed to for serving several countries; if each country decided it needed a regional centre, then the pre-war situation would be re-established and the next stage would be to set up a central transit centre, which represents the first plan again.[10]

5.2.3 *Fundamental Transmission Plan*

Most of the problems in obtaining a high-grade transmission standard will be found in the national areas, particularly where existing plant is inadequate. Between most international *tête-de-ligne* centres there is little doubt that zero-loss, high-velocity carrier systems will be used. The stability of these circuits will be improved by 4-wire switching at the transit exchanges and at the *tête-de-ligne* centres. The conversion from 4-wire to 2-wire will take place inside the national networks. The recommendations of comprehensive studies of international transmission plans have already been published.[11]

It seems likely that most of the circuits required for a transit ring will be of the coaxial type. In this way cables with a large capacity for future growth can be made quickly and

TABLE VI

Forecast for 1952 of Traffic between Transit Centres Arranged in Ring Formation

Clockwise	Brussels	Rotterdam	Copenhagen	Malmö	Danzig	Warsaw	Cracow	Brno	Vienna	Zürich	Paris
Paris	120	36	3	4.5	—	—	—	—	—	—	—
Brussels	—	44	6	3	2.5	—	—	—	—	—	—
Rotterdam	—	—	21	25	8.5	—	—	—	—	—	—
Copenhagen	—	—	—	26	4	—	—	2.3	1	—	—
Malmö	—	—	—	—	7	—	—	2.3	2.3	—	—
Danzig	—	—	—	—	—	—	—	1.1	0.1	0.2	—
Warsaw	—	—	—	—	—	—	—	11	8.5	4.5	6
Cracow	—	—	—	—	—	—	—	17	5	2	—
Brno	2.5	6	—	—	—	—	—	—	20	—	8
Vienna	6	7	—	—	—	—	—	—	—	34	16
Zürich	13	30	3	2	—	—	—	—	—	—	100

Anti-clockwise	Brussels	Paris	Zürich	Vienna	Brno	Cracow	Warsaw	Danzig	Malmö	Copenhagen	Rotterdam
Rotterdam	27	36	30	5	6	—	—	—	—	—	—
Brussels	—	60	14	4	2	—	—	—	—	—	—
Paris	—	—	114	30	8	4	—	—	—	—	—
Zürich	—	—	—	20	4.5	4	—	—	—	—	—
Vienna	—	—	—	—	32	13	—	0.05	1.2	1	—
Brno	—	—	—	—	—	12	—	0.25	2.75	2	—
Cracow	—	—	—	—	—	—	—	—	1.1	0.25	—
Warsaw	4.5	—	—	—	—	—	—	—	4	2.25	6.5
Danzig	—	—	—	—	—	—	—	—	8	2	—
Malmö	2.2	4.1	2.2	—	—	—	—	—	—	24	14
Copenhagen	3.3	3.2	1.2	—	—	—	—	—	—	—	20

Note 1: The traffic is expressed in erlangs (call-hours).

Note 2: The figures assume that traffic to Poland joins the national network at Danzig and Cracow.

Note 3: The traffic shown is not only that from the country in which the transit centre is situated. The clockwise traffic from Zürich includes that from Italy, Yugoslavia, and the Balkans. The anti-clockwise traffic from Zürich includes Spain, Portugal, and Luxembourg.

Note 4: A traffic group greater than 20 erlangs will have a good occupancy, and a group less than 10 erlangs will have a poor occupancy, unless an overflow to an alternative route is available.

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[10] M. Langer, "Studies on Telephone Engineering Problems," R. Oldenbourg, Munich, 1939.

[11] B. H. McCurdy, "Fundamental Transmission Planning of Telephone Networks," *Electrical Communication*, v. 18, pp. 25–32; July, 1939: v. 19, pp. 18–28; July, 1940.

installed at a cost which compares favourably with other means of transmission. Most of the circuits making connection with the ring will also be coaxial, especially where the number of channels required is considerable. In other cases, where the traffic needs only a small number of circuits and older types of cable construction exist, some time may elapse before multichannel transmission systems can be justified.

With coaxial and similar systems, it is convenient if each traffic group can be allocated with a group, or supergroup, of transmission circuits so that they may be switched conveniently without the circuits being terminated individually. If the circuits to a transit centre comprise, for example, 10 different traffic groups, it will be necessary to separate the groups sooner or later, and unnecessary expense can be eliminated if terminating equipment for converting the circuits to audio work is avoided. It is unlikely that the traffic groups will be of such a size that they will exactly fill a supergroup, and in fact it is doubtful whether it will be possible to avoid breaking into groups. Care will be necessary in the planning of the groups to avoid all unnecessary frequency-band transformations. Problems will arise concerning the desirability of combining traffic groups in order to take advantage of reduced quantities in larger bundles. The annual-charge plant expense can be evaluated by a simple formula comparing the length of line with the annual charges. If the line charges (including repeater stations) per kilometre-channel are expressed by C_L, and C_C represents the charges per circuit for switching equipment, C_T the charges for each terminating equipment (including signalling equipment), N_2 is the number of extra channels to be switched, and N_1 the number of physical channels saved, then the length of line must exceed

$$\frac{N_2 \times C_C + 2(N_2 - N_1)C_T}{N_1 \times C_L}.$$

To take an example, let $C_L = 1$, $C_C = 6$ and $C_T = 75$, and let 30 direct circuits be reduced by $6(N_1)$ to $24(N_2)$; then the distance must exceed

$$\frac{24 \times 6 + 2(24 - 6)75}{6 \times 1} = \frac{144 + 2\,700}{6}$$

$$= \frac{2\,844}{6} = 474 \text{ kilometres.}$$

With small groups, it may be found that N_1 exceeds N_2. If the direct-route distance (d_1) differs from the combined-route distance (d_2), then

$$(N_1 + N_2)d_1 \times C_L - N_2 \times d_2 \times C_L$$

must exceed the annual charges

$$N_2 \times C_C + 2(N_2 - N_1)C_T.$$

If the direct route is of different construction and type from the combined route, it may be necessary to use C_{L1} and C_{L2} to represent different charge rates. Similarly, it may be necessary to replace C_T by C_{T1}, C_{T2}, or C_{T3}, where C_{T1} represents the complete termination to audio, C_{T2} represents a simple group filter, and C_{T3} represents only signalling equipment for cases when the through circuit is converted to audio. The annual equipment charge expression would then be

$$N_2 \times C_C + 2N_2 \times T_2 \text{ (or } T_3) - 2N_1 \times T_1.$$

Each switching point added may involve a small additional insertion loss due to the connection of signalling equipment.

5.2.4 Speed of Connection

This subject merits serious consideration in the planning of a new long-distance switching system. An important reason for a revised fundamental plan is to speed up the service. In the past, much of the delay has been associated with the process of operating—not owing to any failing on the part of the operators but because arrangements did not easily permit a high-speed service to be provided. One contributory factor was the inadequacy of circuits, and it is clear that with more automatic switching this hindrance could be removed. It is the general experience that on long-distance circuits the provision of an increased number of working circuits reduces the delays only for a short period, because the better service encourages more traffic until an overload is re-established. It is a common condition to find such traffic growing by 25 per cent per annum in cases where additional channels are provided to maintain a speedy service. The saturation point has not been reached in any country, although the capacity of the subscribers to make long-distance calls must be finite. A second contributory factor has been the result of the many processes involved in establishing an international call, including the preparation of the demand ticket, the transfer to the line position, the request to the distant operator, the recall of the calling party, and the monitoring necessary to ensure that the two parties are making progress with the conversation. The alternative method with semi-automatic operation can be speeded up in the following way:

A. Combined line and recording (c.l.r.) operation.

B. Operator uses keyset, storing required number in local sender.

C. Called number indicated on the combined-line-and-recording operator's position, avoiding the use of pencil and ticket before starting to complete the connection.

D. Connection to register in terminating country established on high-speed basis.

E. Transfer of information, in high-speed code, from local sender to distant register.

F. Preparation of ticket automatically.

With a local sender normally available, the operator can set up the called number immediately it is requested. She can check back with the caller the number set up and therefore recorded on her indicator. It seems probable that the national numbers can be more easily visualized if a combination of figures and letters is used rather than all figures. Compare the following examples:

217	BO2	5 800	and	217	262	5 800
816	HOL	8 765	and	816	405	8 765
94	CE	2 674	and	94	23	2 674

It is assumed that the country required can be selected by the use of an individual key for each country.

The connection between the sender and the terminating register can be achieved very quickly and at little cost if there is a group of direct circuits. Not very much greater expense need be involved if the transit centres are designed to make selection in a few hundred milliseconds each. It seems unlikely that there can be adequate justification for transmitting digits in the ordinary sense for setting up the connection to the register in the country required. For example, a call from London to Milan might possibly use a direct circuit from London to Milan or it might use a circuit from London to Zürich and a new circuit from Zürich to Milan. As a further possibility, it might use London–Paris and Paris–Milan circuits or Paris–Zürich and Zürich–Milan circuits. All different combinations might lie in the same multi-channel transmission system over the complete distance. The sender in London would presumably cause the different circuits to be searched in a definite order to maintain a high occupancy for the direct circuits. Whichever type of circuit is chosen, a particular seizing signal could be transmitted, which would be sufficient to set up the connection to the incoming register at Milan, from which a proceed-to-send signal would be transmitted to permit the national number to be sent. Such a simple and rapid means of completing the connection would hardly be possible with a single centralized transit centre, which would need a numerical signal to choose the correct direction, as the number of possibilities would be much larger. The same difficulty would arise with a number of regional centres all interconnected. A brief seizing signal controlling selection is valuable, not only in time of connection, but also because it tends to reduce the signal time, which needs to be kept as low as possible to avoid overloading the line amplifiers.

The transfer of the call description need not exceed approximately 2 seconds. The selection of the required subscriber in the terminating country may take some seconds because, although the call can be expedited until the local exchange is reached, from this stage the time taken cannot easily be less than for a local call. It seems reasonable that on most international calls the required subscriber's line should be selected within 30 seconds of the request being made. With such a quick service the calling subscriber is likely to be awaiting the call or otherwise capable of being summoned quickly.

Within the next 25 years, it is to be expected that improved automatic systems will be available, permitting relatively high-speed switching of local connections and, unless the international service is designed for high speed, it is liable to be found inadequate and an obstacle to the provision of the highest quality service. The initial cost of the equipment and the maintenance expense are not likely to be altered very much, whether the transit service is handled by quick or by normal switching arrangements.

For these reasons, the initial planning of any new international network should study and strive to incorporate rapidity of operation.

5.2.5 Signalling System

The elementary requirement from the signalling system is simplicity of operation, but experience has shown clearly that this result is not obtained by the use of a simple signalling system which is achieved by limiting the number of different signals. As an example, the ringers used on international circuits represent the most primitive and simple form of signalling and result in difficulties in operating and a low-grade service to the subscribers. To improve and speed up the service, it is necessary to have a universal operating method, an adequate code of signals to allow the operators to work quickly and effectively, signals which are not subject to interference, and a signalling system which promotes, and does not impede, a high-speed service.

The amount of signalling equipment which appears on international circuits is appreciable, but it may very well be a false saving to cut down on this equipment if it means delay and difficulty in operating efficiently. The Comité Consultatif International Téléphonique have studied the question of signalling in some detail, and a number of decisions have been reached.

TABLE VII

SIGNALLING COMPARISON

Objection	Signalling by voice-frequency currents	Signalling over separate channels
Interference from speech	Needs safeguards	None
Interference between systems	Needs safeguards	None
Interference from echo suppressors	Needs safeguards	None
Overloading of amplifiers in multi-channel transmission systems	Needs safeguards	Reduced liability to trouble
Speed of signalling:— (*a*) Direct (*b*) Transit	 High High	 High Needs delay for repetition
Passage of signals through transit exchange	Automatic	All signals must be transferred to direct current and superimposed on talking circuit
Line expense	None	Variable. Probably minor if quantities are large
Possibility of establishing connection on channel on which speech circuit is out of order	Unlikely	Possible
Distortion due to repetitions	Very small	Increases with each repetition

As on most international circuits, it is to be expected that 4-wire high-velocity speech circuits will be used. The use of direct-current or 50-cycle-per-second alternating-current signalling systems is ruled out, because such signals will not be transmitted over the carrier circuits. The interruption of the transmitted carrier current for signalling purposes is undesirable on account of intermodulation effects. The two main alternatives which remain are the use of voice-frequency and separate signalling channels; opinion favours the former. The relative advantages and disadvantages of the two methods are shown in Table VII.

It can be seen that most of the objections to voice-frequency systems concern interference. Interference from speech currents is liable to occur, especially if the signalling power used is very small and the receivers are very sensitive. It has been found that guard circuits provide a satisfactory safeguard, especially if the signal length is not too short. Most of the circuits equipped with voice-frequency apparatus in Europe have used either 600 and 750 cycles per second or 2400 and 2500 cycles per second. It is claimed that greater speech immunity can be obtained by using 600 and 750 cycles per second simultaneously, and tests carried out indicate that for any fixed time-delay the interference with the single frequencies is several hundred times that with the frequencies used simultaneously. This fault liability can be balanced by increasing the signal length, 150 milliseconds of compound-frequency being roughly equivalent to 400 milliseconds of simple-frequency operation. It seems likely that other frequencies, such as 2000, 2400, 2700 and 3000 cycles per second, may also be superior to 600 and 750 cycles per second, and tests are being undertaken to establish to what extent this is true. The frequency/response efficiencies of the subscribers' transmitters and the cut-off frequencies of existing cables are other aspects that must be considered at the same time. The existing recommendation by the Comité Consultatif International Téléphonique is to use 600 and 750 cycles per second, and the frequency or frequencies eventually chosen will have little bearing on the speed of the service and not a great effect on the expense of the equipment involved.

TABLE VIII

COMPARISON OF SELECTIONS OF NATIONAL SERVICE-TONES

Tone	Austria	Denmark	France	Germany	Great Britain	Sweden	Switzerland
Dialling	Uninterrupted		Interrupted in cadence (66/33 c/s)	Uninterrupted	Uninterrupted (33 c/s)	Uninterrupted	Uninterrupted
Ringing	1-sec tone 5-sec silence	1-sec tone 3.5-sec silence	1.66-sec tone 3.33-sec silence	1-sec tone 5-sec silence	0.4-sec tone 0.2-sec silence 0.4-sec tone 2-sec silence	1-sec tone 9-sec silence	1-sec tone 5-sec silence
Busy (normal)	0.5-sec tone 0.2-sec silence 0.5-sec tone 1-sec silence	0.175-sec tone 0.47-sec silence	0.5-sec tone 0.5-sec silence	0.5-sec tone 0.2-sec silence 0.5-sec tone 1-sec silence	0.75-sec tone 0.75-sec silence	0.25-sec tone 0.25-sec silence	0.25–0.50-sec tone 0.25–0.50-sec silence
Busy (subscriber busy but "no call waiting" given to operators only)			0.5-sec tone 2-sec silence				
Number unobtainable					5-sec tone 1-sec silence	0.25-sec tone 0.25-sec silence 0.75-sec tone 0.25-sec silence	0.5-sec tone 0.1-sec silence 0.1-sec tone 0.1-sec silence 0.1-sec tone 0.1-sec silence
Information		0.06-sec tone 0.09-sec silence 0.06-sec tone 0.09-sec silence etc., every 10th silence of 0.6-sec duration					

The possibility of interference to and from other systems has not been sufficiently widely understood; the trouble is due to signals passing naturally from end to end of a connection, and, in consequence, signals transmitted by one system may cause an unwanted response in another system. This can result in an incorrect signal or irregular release and therefore needs safeguarding. The Comité Consultatif International Téléphonique have recommended the use of a prefix signal which opens the line, thereby isolating the essential portion of the signal and preventing it from passing out to cause interference in any other system. This safeguard may delay the answer signal by 50 milliseconds and cause a small increase in cost, but it involves no real technical difficulty.

Interference from echo suppressors is a more serious menace to voice-frequency systems. One method of avoiding such interference is to repeat the signals until they have been acknowledged. This process adds to the stability of the service but it increases the cost and complication of the equipment. As future international circuits will be almost entirely of the high-velocity-carrier type, the trouble is only temporary, and it may be found possible to eliminate the interference by arranging the position of the echo suppressors so that they are not connected between the voice-frequency receivers.

The overloading of the line amplifiers in multichannel transmission systems is probably the most difficult problem. These amplifiers have sufficient capacity for normal speech currents, but a large proportion of the conversation contains no speech at all in one direction or the other, so that the mean energy level is low and the permissible power and signal duration should both be limited.[12–14]

5.2.6 *Service Tones*

The difference in national service-tones is a good example of the difficulty which can arise as a result of independent development by separate administrations. At the present time,

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[12] W. G. Radley, and E. P. G. Wright, "Voice Frequency Signalling and Dialling in Long-Distance Telephony," *Journal I.E.E.*, v. 89, Part III, p. 43; 1942.

[13] F. P. O'Grady, "Dialling on Long Distance Trunk Lines—A Review of Developments 1944," *Telecommunication Journal of Australia*, Feb., 1945.

[14] H. Jacot, "Développement de la signalisation et la sélection automatique à fréquences vocales dans le réseau téléphonique suisse," *Bulletin de l'Association Suisse des Electriciens*, No. 21, 1946.

there are so many different codes of service tones in operation that it would be a real difficulty for the originating international operator to recognize the significance of any particular tone.

The recommendation made by the Comité Consultatif International Téléphonique in 1938 is too general to assist recognition, and administrations have been reluctant to make widespread changes to achieve any single standard. Table VIII shows a comparison of some national service tones, and it is hoped that it may be possible to overcome the difficulty mentioned in the following way:—

Dialling Tone	This will be generated in the adjacent sender. The voice-frequency proceed-to-send signal can be standardized. It will indicate when transmission of the stored information can be commenced.
Ringing Tone	An effort will be made to identify this tone in the incoming *tête-de-ligne* centre. On recognition, the line will be opened and a standard voice-frequency signal transmitted which will enable the outgoing *tête-de-ligne* centre to connect the national ringing tone, which would be recognized by the caller. When the call is answered the line is reconnected.
Busy Tone (Transit Centre)	This can easily be turned into a voice-frequency signal and enable the outgoing *tête-de-ligne* centre to connect the national busy-tone and clear the international line.
Busy Tone (Subscriber)	An effort will be made to identify this tone in a similar way to the ringing tone, and to apply a standard voice-frequency signal which will enable the national busy-tone to be connected.
Miscellaneous, Number Unobtainable and Changed-Number Tones	It will be necessary for the outgoing operator to seek assistance from an incoming operator in the event of receiving an unrecognized tone. As most of the tones of this type need operator assistance this procedure may not cause any great delay.

It seems that this solution does not go far enough, and further study should be undertaken to investigate the possibility of reducing the present non-uniformity by agreeing upon a standard code which should become operative in, say, 20 years' time.

Uniformity of tones is valuable on national calls as well as international calls, because many travellers abroad are handicapped in making calls on account of the difficulty of recognizing the service tones.

5.2.7 Language Differences

International calls fall into two general classes, one in which the caller is intending to speak in the language of the country which he is calling, and the other in which the caller is intending to speak his own language and may not understand the language of the country which he is calling.

In the former case, especially if the traveller is speaking to his home country, most of the difficulty is experienced in obtaining access to the international operator and passing the details of the call. Once he is connected to his own country the only language difficulty will be that of the international operator in trying to determine whether the required connection has commenced. This should be indicated by the supervisory signals, and it is essential that the code of signals should make full provision for giving complete supervision.

If the caller is making a call to an individual in a foreign country and is unable to converse in that language, it is probable that he will originate a personal call to ensure reaching someone with whom he can talk. The line operator who is unable to speak the language concerned will presumably transfer such calls to an operator speaking the language or else call the inward *tête-de-ligne* operator to obtain her assistance.

TABLE IX

LANGUAGE DIVISIONS

Belgium	2 main languages, Flemish in north, and Walloon in south
Bulgaria	1 main language
Canada	2 main languages
China	Many variations
Czechoslovakia	2 main languages
Eire	2 main languages
Finland	1 main language, 2 fringe districts differ
Great Britain	1 main language, 1 variation
Greece	1 main language
Hungary	1 main language, 3 fringe variations
India	5 main languages, with variations
Poland	1 main language, 3 fringe variations
Rumania	2 main languages, 1 fringe variation
Switzerland	3 main languages
U.S.S.R.	Several variations
Yugoslavia	3 main languages, 4 fringe variations

Note: In addition to the variations in language, most countries have one or more dialects.

Language problems are likely to be complicated by the fact that several countries are bilingual while in other countries there is a considerable variation in dialect. Table IX emphasizes these variations.

It is abundantly clear that in spite of every precaution being taken at the originating *tête-de-ligne* centre there will be cases when an unexpected language is encountered in setting up an international call, and it is unreasonable that these calls should be broken down and re-set up through a second operator. Such a practice would tend to encourage the outward operator to set up all calls in this way in order to ease the operating. What is required is a signal that will enable the outward international operator to summon rapidly an inward international operator. It is believed that a forward transfer signal of this character could be incorporated without undue difficulty, and that such a signal is, in fact, used in some national voice-frequency signalling systems. One important rule which the language difficulty emphasizes is that the outward international *tête-de-ligne* operator should supervise the connection; any intermediate zone-centre operator who has helped to extend the call to the international centre should pass the control forward.

5.2.8 Tariff

Every administration has a settled scale for regular and special-attention calls, and the new toll plan does not need any fundamental change to the tariff structure. The record of the call can be made on a ticket prepared either by the international operator or by a machine; possibly in the former case, but certainly in the latter, the charge calculation will be automatic. In order that the operator may concentrate on setting up the connection, there are obvious advantages in allowing the ticket to be printed independently, and it is clearly desirable to employ some form of machine ticketing with any application involving subscriber dialling. In addition to the charge calculation, it is necessary to inform the caller, on request, of the charge to be made. It would be possible to provide "duration and charge advice" positions on which the details of the calling and called numbers, the duration of the call and the charge were all indicated; this information could then be passed to subscribers without delaying the long-distance operators.

It has recently been agreed that the international accounting shall be based on the tickets made at the originating centres. Such an arrangement is most appropriate for a high-speed semi-automatic or high-speed full-automatic service. If it is necessary for the transit or terminating countries to obtain information about the identity of the calling subscriber, more equipment is needed to record the information, and there is consequently a danger that the service may be retarded. The decision to use the ticket from the originating country need not deprive the transit and terminating switching centres of necessary information. There is a need for some method of checking traffic flow in order to determine the seasonal variations, the normal rate of growth and how much traffic is directed to different countries. It is clearly more difficult to collect information on these subjects on international than on national circuits. With the plan as outlined, it would be relatively simple to record at the transit centres the route on which a call occurs, the route to which it is switched, and the duration of the connection. These details could be recorded by ticket machines and the information analysed to ascertain the traffic characteristics.

5.2.9 Alternative Routes

Another subject associated with tariff rates, which is not easy to evaluate, is the best method of handling alternative routes. It should be made clear that emergency routes are subject to special-rate regulations which are not expected to cause serious complications in the future.

The alternative-route problem should be divided into two categories, since two quite separate principles are involved. In the first place it has been an operating practice for many years to have alternative means for a switchboard at A to reach a remote switchboard at B via a switchboard at either C or D, and sometimes through other routes as well. These diversions are based on the non-coincidence of busy-hour traffic. From the operating standpoint it is rather more elementary; if A finds that all the channels to C are in use she applies to D for a channel. If *D* has no direct channel to *B*, then *D* may ask *E* for a channel to *B*. The call is established, and it is difficult to know how much traffic is displaced by the diversions. If the groups are large, very little is gained; but if not, there is a serious tendency for traffic from a large group to swamp a small group. The arrangement as a whole is unsound, except as a temporary expedient.

The second application is based on a different principle. In this case certain direct routes are calculated on a high-loss rate, such as one call in four, and the overflow is arranged to be carried over transit groups which are materially larger than the direct groups. The transit groups are designed to cater for this overflow and are, in fact, less liable to variation in busy-hour traffic than individual groups. This arrangement has been used widely both in America and in Europe; it leads to considerable economy on direct circuits at very little cost on the transit circuits. On international circuits, the indirect route is liable to pass through an additional country. It is not difficult to record the traffic which uses the alternative route and to credit the third administration, so that the essential requirements can be met. There remains the fact that, on perhaps 5 per cent of the calls, a third administration draws a fee to the detriment of the administration in whose territory the call originated, and it is possible that the thought that something is being lost will be misunderstood and that an effort will be made to provide more direct circuits on an uneconomical basis, resulting in increased annual charges on the direct circuit. In this way, a greater loss would be incurred than by obtaining assistance from a third administration. In many cases reciprocal action would tend to balance the account, but it is certainly doubtful whether the basic economics of the arrangement will prevail where several administrations are concerned.

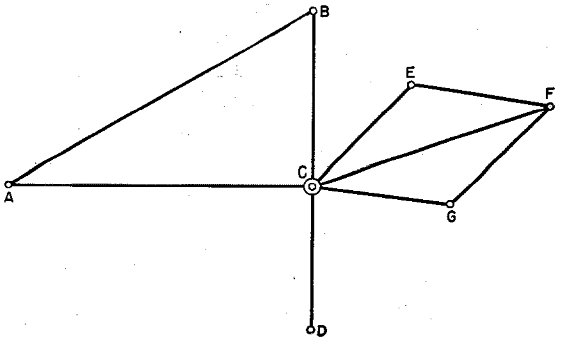

Fig. 3—Alternative routing with one principal transit centre.

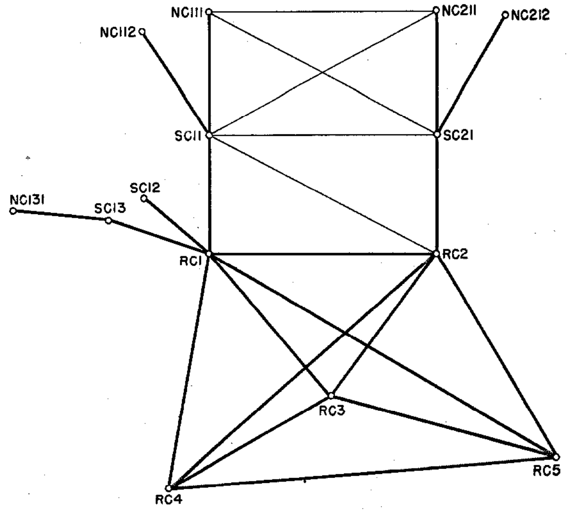

Fig. 4—Alternative routing with interconnected regional centres.

— High probability of loss circuits.
━ Low probability of loss circuits.

Figs. 3, 4, and 5 show the differences in the three cable layouts already considered as regards alternative routes.

In Fig. 3, *A* has a direct circuit to *B* and can overflow reasonably through *C*, unless *A–B* traffic is greater than *A–C* or *B–C* traffic. If *A* normally reaches *F* through *C*, and *A–C* is engaged, an attempt can be made through *B*, but this is satisfactory only if *A–B* and *B–C* are greater than *A–C*. Similarly, if *C–F* is engaged, overflow may be effected through *C–G* and *G–F* or *C–E* and *E–F*, provided these routes are greater than *C–F*, which is doubtful. It will be seen that the first case is opposed to the second. The efficiency of the plan depends very much on the length of *A–C*.

In Fig. 4, the basic plan presupposes that the regional centres *RC* are all fully interconnected and that each *RC* has a number of section centres *SC* to which, in turn, are connected national centres *NC*. All these circuits are high-grade, with a normal probability of loss. In addition, direct circuits with a high probability of loss may connect adjacent centres. In this case, a call from *NC*111 to *NC*211 might make a primary attempt at a direct connection, or, if this were not possible, an attempt through *SC*21, it being fundamental that *SC*21 has a normal-loss circuit to *NC*211. If this failed, *NC*111 would pass the call to *SC*11, which might try a high-loss circuit to *NC*211 and, if it had the direct circuits, via *SC*21 or *RC*2. Failing to make connection, the call is passed to *RC*1, and from there to *RC*2, *SC*21, and *NC*211 on normal-loss circuits throughout. In such an arrangement a high efficiency of direct circuits is obtained, which is very desirable if the geographical situation of *RC*1, *RC*2, etc., tends to involve heavy back-hauls. It should be noted that the concentration of the *RC* transit centres adjacent to one another is purely to simplify the sketch; in practice these transit centres would be well dispersed over the whole area.

In Fig. 5, *A* may have direct circuits to *K*, *L*, and *M* at a high probability of loss; also overflow access through the ring, which would provide normal-loss circuits. Alternative routes would also be available from *A* to *N* by means of *E* or *F*, which might be reached in the same direction of traffic or in opposite directions around the ring. If the international operator sets up the call requirements into adjacent senders it is probable that any alternative routing to be introduced will be independent of any special action of the operator. This is correct, as it is undesirable to burden the operator with considerations of which routes should be tried or, in fact, to allow any possibility of the operator setting up unauthorized routes.

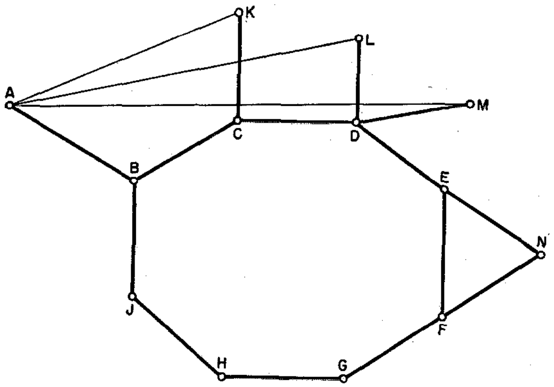

Fig. 5—Alternative routing with ring formation.

High probability of loss circuits.
Low probability of loss circuits.

The quantities shown in Table V are based on the use of alternative routes, and it is noteworthy that a considerable economy of circuits is made possible. This can be appreciated from Table X,

TABLE X

COMPARISON OF CABLE ROUTES EXPRESSED IN CIRCUIT-KILOMETRES WITH AND WITHOUT ALTERNATIVE ROUTES

Formation of cable plan	Type of circuit	*With* alternative routes	*Without* alternative routes
Star (Fig. 3)	Direct	156 350	167 120
	To centre	8 000	10 000
	Terminations (overflow)	9 450	—
	Terminations (indirect)	6 430	25 800
		180 230	202 920
Ring (Fig. 5)	Direct	65 670	66 330
	Ring	33 040	37 500
	Terminations	75 200	79 180
		173 910	183 010

which compares the layouts (for Austria) with and without alternative routes for the star and ring formations corresponding to Figs. 3 and 5.

The quantities in Table V demonstrate that the star arrangement is more dependent on alternative routing than is the ring arrangement, except in the case of Germany. With the ring formation, the direct and alternative routes are frequently passing through the same countries, but with the star formation it is usual for the direct and alternative routes to pass through different countries.

5.2.10 Numbering Plans

In comparison with North America, the switching in Europe, following different practices, is necessarily divided into more isolated groups. There is likely to be little desire for a continental numbering plan, and therefore the problem of numbering becomes a national affair, with a simple addition for the country selection. Both in America and in Europe, it is likely that there will be cases where a single numbering plan will spread over a national border, e.g., Buffalo, Detroit, etc.

5.3 SWITCHING EQUIPMENT

The desire to set up connections rapidly would appear at first sight to set some restrictions on the type of switching equipment used in the transit centres, but it is by no means certain that high speeds cannot be obtained with all the principal systems by some appropriate circuit rearrangement. There is also an argument that standardization is necessary to enable all transit centres to provide the necessary facilities. On investigation it seems that the argument is not as powerful in this case as in that of signalling. Non-standardization in signalling means repetition, involving additional equipment and delay. On the other hand, there is a good argument that the equipment to be used at the transit centres should be similar to that used normally for national service. Similarity in this way will ensure adequate maintenance knowledge and frequently mean that additional equipment can be obtained more quickly.

TABLE XI

TYPES OF SWITCHING SYSTEM

Algiers, Morocco and Tunisia	R.6 and French step-by-step
Argentina	English step-by-step
Australia	English step-by-step
Austria	Austrian step-by-step
Belgium	Chiefly rotary, some step-by-step
Brazil	Rotary, American and English step-by-step
Bulgaria	German step-by-step
Canada	Canadian step-by-step
Chile	English step-by-step
China	English and German step-by-step
Cuba	American step-by-step
Czechoslovakia	Chiefly German step-by-step, some rotary
Denmark	Rotary, some Swedish Ericsson
Eire	English step-by-step
Finland	Chiefly German step-by-step, some Swedish Ericsson, and some crossbar
France	Paris and other big cities: rotary. Provinces: R.6
Germany	German step-by-step
Great Britain	5 cities: director type step-by-step. Provinces: normal step-by-step
Greece	German step-by-step
Hungary	Rotary
India	English and German step-by-step
Italy	German step-by-step, Swedish Ericsson and rotary
Japan	English, German, and Japanese step-by-step
Netherlands	Chiefly German step-by-step, some rotary, and Swedish Ericsson
New Zealand	English step-by-step, rotary
Norway	Rotary, some Swedish Ericsson
Poland	Chiefly Swedish Ericsson, some English step-by-step, and some rotary
Portugal	English step-by-step
Rumania	Rotary
Russia	Swedish Ericsson
South Africa	English and German step-by-step
Spain	Rotary
Sweden	Swedish Ericsson
Switzerland	Chiefly Haslar, some rotary, and some German step-by-step
Turkey	Swedish Ericsson and rotary
U.S.A.	Crosbar, panel, and American step-by-step
U.S.S.R.	Swedish Ericsson
Yugoslavia	Swedish Ericsson and rotary

The variations in the use of systems in different countries are very striking, as can be seen from Table XI.

It can be imagined that there might be some difficulty in deciding which system should be made a standard for transit centres.

5.4 SUBSCRIBER DIALLING

This paper has referred chiefly to the provision of a semiautomatic service, but it is considered that as far as possible the plans should be suitable for subscriber dialling also. Objections are raised to the idea of subscriber dialling, but, from experience with long-distance dialling in national systems, many of the objections do not seem justified.

The case for international subscriber-dialling is as follows. Although many calls will continue to need assistance from an operator, there are many which the subscribers would prefer to make themselves. Some calls are little longer than local-area calls, e.g., Detroit and Windsor, Antwerp and Rotterdam, etc.

The objections to subscriber dialling are that most international calls are relatively expensive and that the subscriber has a right to the assistance of an operator, that servants and small boys will be free to set up long-distance and expensive calls, and that in the event of a wrong number the subscriber will be charged a considerable sum which he will resent.

The response to these arguments is that if the subscriber prefers to dial some calls direct, why should he be forced to have the assistance of the operator? Also, servants and small boys can ask for international calls through an operator; and it is not necessary to charge a 3-minute fee for a call which is released in under 15 seconds; it is found on long-distance national dialling that 10 seconds is sufficient to recognize a wrong number, and the charge for this period has been made the same as for a local call.

Two practical difficulties to subscriber dialling are the need for a dial number-plate carrying letters and figures, and the large expense involved in installing identification circuits for all subscribers where they do not exist already for the national service. It seems a possibility that an international dialling service may commence as a limited service, restricted to certain lines equipped for such a service. It seems very likely that if such a service were made available, a very high proportion of the requests would come from subscribers with private branch exchanges, operated by qualified operators.

6. *National Telephone Systems*

Appendix, Section 8, contains technical information of general interest concerning the national systems in Argentina, Belgium, Denmark, France, Hungary, The Netherlands, Portugal, Sweden, and Switzerland.

7. *Acknowledgments*

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8. *Appendices*

8.1 TELEPHONE INFORMATION ABOUT ARGENTINA

8.1.1 Toll Traffic

TABLE XII

NATIONAL AND INTERNATIONAL TOLL TRAFFIC IN MILLIONS OF CALLS

	As estimated in 1937	Actual
1930	—	7.5
1932	—	7.2
1934	—	7.8
1936	—	10.0
1938	13.0	12.7
1940	15.4	13.9
1942	17.7	16.0
1944	20.3	18.7
1945	21.8	20.0

8-year increase, 1930–1938 = 70 per cent.
7-year increase, 1938–1945 = 57.5 per cent.

8.1.2 International Traffic

The outgoing international calls in 1945 were 200 000, an increase of 87 per cent on 1938; the incoming international calls were 212 000 in 1945, an increase of 56 per cent on 1938.

Note: In Chile, the outgoing international calls increased between 1940 and 1945 by 233 per cent, and the incoming calls by 200 per cent. A large percentage of the international traffic from Chile passes into Argentina.

8.1.3 Primary Toll Centres

Buenos Aires
La Plata
Mar del Plata
Necochea
Tandil
T. Arroyos
Bahia Blanca
Pehuajo
Chililcoy
General Pico
Junin
Rufino
Rio IV
Pergamino
Va. Maria
Cordoba
Rosario
Santa Fé
Reconquista

Buenos Aires will have direct circuits to all other centres, except possibly Reconquista. The mean length of these circuits is approximately 400 kilometres, and the maximum about 800 kilometres. Buenos Aires will also have tributary circuits to another 45 centres with distances varying between 50 and 300 kilometres.

Other long-distance lines radiate to Mendoza, Tucuman, and Entre Rios. Land lines provide international circuits to Santiago and Montevideo.

There are radio connections to New York, a combined circuit for Madrid, Paris, London, and Berlin, and independent circuits to Rio de Janeiro, Bariloche, Lima, and La Paz.

8.1.4 Long-Distance Dialling

Thirty of the circuits between La Plata and Buenos Aires are arranged for operator key-sending, using a register at La Plata and a 50-cycle-per-second current for passing the information over the toll line.

The circuits between Mar del Plata and Buenos Aires operating over 384 kilometres employ direct-current dialling. These circuits are open wire and subject to low-resistance leakage to ground due to heavy mist and flying cobwebs.

The long-distance circuits will be converted from manual to semi-automatic working in accordance with the fundamental plan.

8.1.5 Transmission Systems

The country is suitable for carrier circuits, and it is expected that a considerable extension of 12-channel and 3-channel systems will be brought into service in the next 5 years.

8.1.6 National Numbering Scheme

No definite plan is yet established, but the ultimate nation-wide operator-dialling layout is likely to incorporate an overall numbering scheme.

8.1.7 Manual Toll-Traffic Switching

The present distribution of toll traffic by switching points is approximately:

Direct, 80 per cent
1 switch, 16.5 per cent
2 switches, 3 per cent
3 or more switches, 0.5 per cent,

the operating time-factor being:

Direct 1.0
1 switch, 3.14
2 switches, 5.28
3 switches 8.14

8.1.8 Relationship between Distance, Duration, and Percentage of Revenue for National Long-Distance Calls (See Table XII)

8.2 Telephone Information about Belgium

8.2.1 Transmission

Two-wire lines are used so long as no more than two toll-line sections are connected in tandem. These 2-wire lines are divided into separate groups for terminal and tandem traffic.

Four-wire lines are used for all except the first and last sections when the number of sections connected in tandem exceeds two.

8.2.2 Routing

If a secondary exchange is connected to two primary exchanges, both incoming and outgoing traffic has to be routed over the shorter route.

8.2.3 Tandem Routing

Up to 5 primary toll exchanges may be involved in one connection.

8.2.4 Numbering

The country is divided into 47 zones, each of which has a closed numbering system. Two or three zones have, or will have, 6-digit numbers and will be reached by a 2-digit toll prefix; the remaining zones have 5-digit local numbers and will be reached by a 3-digit toll prefix. The total number of digits to be dialled for calls outside the local zone is therefore 8 in every case.

8.2.5 Metering

All toll charges, with the exception of those for calls to neighbouring exchanges within the local call zone, will be registered by automatic ticket printers. The tariff is determined by the distances between the centres of each "sector," 23 of the so-called "larger zones" being divided into a number of "sectors," which may vary between two and six, whereas the remaining small zones have only one "sector." The fee may be determined by the prefix and the two or three first figures of the subscriber's number.

TABLE XIII

Distance kilometres	Mean duration minutes	Percentage of total calls	Percentage of total revenue
0–50	2.4	41.72	12.15
51–100	2.73	26.84	14.93
101–150	2.91	7.81	7.71
151–200	3.1	4.93	6.68
201–250	3.24	4.92	9.12
251–300	3.35	4.62	11.2
301–350	3.45	1.73	4.56
351–400	3.55	3.27	10.34
401–450	3.61	0.76	2.65
451–500	3.68	0.76	2.97
501–600	3.75	0.92	4.47
601–700	3.81	0.904	4.96
701–800	3.88	0.139	0.97
801–900	3.95	0.042	0.34
901–1 000	4.01	0.353	3.62
1 001–1 100	4.08	0.163	1.77
1 101–1 200	4.15	0.003	0.04
1 201–1 300	4.21	0.1	1.34
1 301–1 400	4.28	0.014	0.16
1 401–1 500	4.35	0.002	0.02

8.2.6 Time Limit

Forced release takes place after 99 minutes of conversation, but means are provided to release in case of emergency after 3 minutes.

8.2.7 Traffic Metering

A traffic meter is provided in each toll register in order to indicate the total number of connections set up to each zone. All remaining traffic information is obtained from the tickets prepared by the automatic ticket printers.

8.2.8 Registers

Arranged for a total of 8 figures (as explained in Section 8.2.4), code translation for 1 to 6 selections of 1 to 20 impulses, and for controlled action with "seize" impulse and "end-of-selection" impulse. Impulses are transmitted in the forward direction.

8.2.9 Outlet Searching

All toll lines work in ideal groups.

8.2.10 Selection Time

No maximum fixed.

8.2.11 Alternative Routing

Provision is made for alternative routing independent of register translation.

8.2.12 Impulsing

Primary toll centres are provided with impulse correctors.

8.2.13 Signalling

By 50-cycle-per-second current; also on 4-wire lines.

8.2.14 Signals

The same as for Switzerland (see Section 8.10), except for "offering" which is omitted.

8.2.15 Pulse Release

Subscriber dialling, 30–60 seconds.
Incoming register, 30–60 seconds.
Incoming tandem line, 5–10 seconds.
Busy or ringing condition—operator is called in after 30–60 seconds.

8.2.16 Insertion of Repeaters

Each incoming tandem line is provided with a repeater, cancelling the loss of the associated line.

8.2.17 4-Wire Switching

This is provided.

8.2.18 Prefix for International Traffic

004.

8.3 Telephone Information about Denmark

8.3.1 National Numbering Scheme (See Table XIV.)

TABLE XIV

*01	Copenhagen	†055	Nyköbing Fl.
		056	Maribo
021	Helsingör	057	Nakskov
022	Hilleröd	058	Vordingborg
023	Frederiksund	059	Stege
024	Roskilde		
025	Holboek	*061	Aarhus
026	Kalundborg	062	Randers
027	Köge	063	Grenaa
028	Ringsted	064	Rönde
029	Noestved	065	Odder
020	Rönne	066	Horsens
		067	Skanderborg
†031	Odense	068	Silkeborg
032	Svendborg	069	Hammel
033	Faaborg	060	Tranebjoerg
034	Assens		
035	Middelfart	†071	Herning
036	Bogense	072	Skjern
037	Kerteminde	073	Ringköbing
038	Nyborg	074	Holstebro
039	Rudköbing		
		†075	Skive
*041	Kolding	076	Viborg
042	Vejle	077	Lemvig
043	Fredericia	078	Hurup
044	Sönderborg	079	Thisted
045	Haderslev	070	Nyköbing M.
046	Aabenraa		
047	Tönder	†081	Aalborg
048	Gram	082	Skagen
049	Holsted By	083	Frederikshavn
		084	Hadsund
†051	Esbjerg	085	Hobro
052	Varde	086	Aars
053	Grindsted	087	Fjerritslev
054	Ribe	088	Brönderslev
		089	Hjörring

* International *tête-de-ligne* centre.
† National transit centre.
Remainder are group centres.

8.3.2 Local Numbering Scheme

Each of the 67 groups has an independent closed numbering scheme. For the smaller groups, it is a 4-digit system; some larger groups are on a 5-digit basis, while Copenhagen is 6-digit.

8.3.3 Alternative Routing

This is anticipated.

8.3.4 Signalling Frequencies

50 cycles per second and 2400/2500 cycles per second.

8.3.5 Signals in Use or Proposed

- Seizing signal
- Impulsing signal
- Answer signal
- Clear-back signal
- Clear-forward signal
- Proceed-to-send signal
- Backward end-of-impulsing signal
- Ring-forward signal
- Trunk-offering signal
- Breakdown signal
- Release signal

8.3.6 Number of Subscribers' Lines

City subscribers	342 000
Rural subscribers	105 800
Official lines	2 700
	450 500

Instruments, including 2- and 4-party lines—558 150.

8.3.7 Operating Administrations

State	Möen South Jutland Toll network beyond 130 kilometres
Copenhagen Telephone Company	Zealand Isle of Bornholm
Lolland Falster Telephone Company	Lolland and Falster
Funen Telephone Company	Funen
Jutland Telephone Company	North Jutland
Saux Telephone Company	Saux

8.3.8

Copenhagen is semi-automatic; Odense has a call distribution system, and most of the other cities are manually operated.

8.3.9

Rural areas are 99 per cent manual. There is a total of about 1500 exchanges. Haderslev (2400) and Aabenraa (1800) are centres of rural automatic areas in South Jutland.

8.3.10

The new toll board at Aalborg will be of the cordless type.

8.3.11

A 12-channel carrier system is used between Copenhagen and Aalborg. There will probably be both 4-wire and 2-wire switching, the latter arranged for 4-wire switching, ultimately.

8.3.12 Tariffs

Local call	5 öre.
Less than 10 kilometres	10 öre.
Between 10 and 15 kilometres	15 öre.
Longer distances	Up to 40 öre.

8.4 Telephone Information about France

8.4.1 Transit Centres

Principal: Paris, Lyons, Marseilles, Bordeaux, and Strasbourg; all interconnected by 4-wire circuits.

Ordinary: Rouen, Lille, Amiens, Nancy, etc. Each is connected to at least two of the principal transit centres.

8.4.2 Distribution Centres

Calais, Dunkerque, Dieppe, etc. Each is connected to one, or several, transit centres.

8.4.3 Trunk and Toll Circuits

These are divided into two groups and four categories:

Group 1.—Transit circuits.
Group 2.—Terminal circuits.

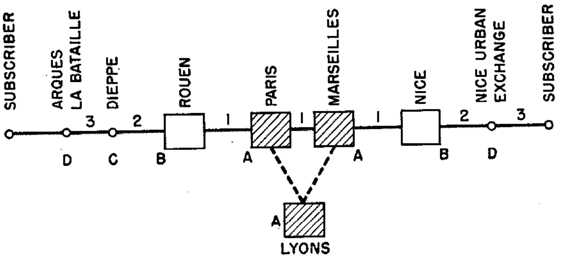
Category 1—4-wire circuits interconnecting transit centres.

Category 2—2-wire circuits interconnecting transit centre and distribution centre.

Category 3—Local junctions and subscriber circuits.

Category 4—Terminating circuits.

8.4.4 *Example of Built-Up Connection*



A—Principal transit centre. **C—Distribution centre.**
B—Ordinary transit centre. **D—Group centre.**

8.5 GENERAL PROPOSALS FOR FUTURE NATIONAL NUMBERING SCHEME

8.5.1 *National Numbering*

All national numbers will be on an 8-digit basis. Subscribers in Paris and the Seine Department will have a 7-digit regional number prefixed by a single digit to form the national number. All other subscribers will have a 6-digit regional number coupled with a 2-digit prefix.

The existing Paris numbers will be modified from a regional to a national basis by the addition of the prefix "1," e.g., VAU 2570 will become 1 VAU 2570.

Existing numbers in Lyons and Marseilles including a letter and four digits will add one letter to complete the regional number and two digits for the national prefix, e.g., (Lyons) F(ranklin) 4322 will become 53 FR 4322.

Existing 5-digit numbers will add a letter for the regional number and two digits for the national prefix, e.g., Rouen 25381 will become 27 R2 5381.

Existing 4-digit numbers will add a letter and a digit for the regional number and two digits for the national prefix, e.g., Elboeuf[15] 2345 will become 27 R8 2345.

8.5.2 *Telephone Regions*

The telephone region covers a district with a local community of interest, and the 6-digit regional number suffices for all connections in this district. Each region may contain a number of towns in different departments, e.g., the Lyons region would include St. Etienne, Grenoble, Annecy, Chambery, Bourg (i.e., the whole of the Department of the Rhône), and a part of the Ain, the Haute-Savoie, the Savoie, the Isère, and the Loire.

To simplify the dialling and the taxation, each department may be given a characteristic national prefix, in which case the telephone region required may be identifiable only with the first digit of the regional number.

8.5.3 *Long-Distance Prefix*

A subscriber obtains a long-distance call by the general prefix "16." Hence a Lille subscriber would call Paris, Lyons, and Rouen as follows:—

16	1	VAU 2570
16	53	FR 4322
16	27	R2 5381

8.5.4 *Metering*

A. Calls within canton	Single fee.
B. Calls within department or between adjacent departments	Determined by distance between chief places in cantons.
C. Calls between non-adjacent departments	Determined by distance between chief towns in departments.

8.5.5 *Time Limit*

Not determined.

8.5.6 *Register and Translator Functions*

The facilities provided with the R6 system for exchanges outside Paris, Marseilles, Nantes, and Arques would be as follows:—

A. Regional calls	Translation for three selections in addition to the 1st selector. Retransmission of part, or all, of the six digits received.

---

[15] Elboeuf is a satellite of Rouen.

B. National calls Receives "16" and eight digits without requiring a second dialling tone. Retransmits the eight digits to district national register.

The register sends a number of impulses not exceeding 18 to a meter-control circuit which operates the subscriber's message register every three minutes after the conversation has commenced.

8.5.7 Selection Time

2–5 seconds.

8.5.8 Alternative Routes

These will be provided.

8.5.9 Impulsing

Impulse correctors will be used.

8.5.10 Signalling

Tests are being carried out with 50 cycles per second, 600 and 750 cycles per second compound, and 2000 cycles per second.

8.5.11 Code of Signals

A. Terminal traffic:—
Seize, proceed-to-send, impulsing, answer, release.

B. Transit traffic:—
All terminal traffic signals, forward end-of-selection, backward end-of-selection, called-subscriber-free signal, clearback signal, backward-recall signal.

TABLE XV

Name	Number of Lines	Name	Number of Lines
Budapest	80 000	Balassagyarmat	1 250
Hatvan	1 250	Szolnok	2 500
Kecskemét	1 500	Székesfehérvár	1 750
Komárom	1 200	Miskolc	2 750
Eger	1 100	Sátoraljaujhely	600
Debrecen	2 800	Nyiregyháza	1 900
Mátészalka	600	Karcag	700
Békéscsaba	2 500	Széged	3 800
Szekszárd	1 200	Baja	1 600
Pécs	2 800	Kaposvár	1 300
Dombóvár	600	Barcs	500
Balatonföldvár	600	Keszthely	700
Nagykanizsa	1 100	Zalaegerszeg	700
Pápa	750	Veszprém	1 100
Szombathely	1 600	Györ	2 200
Sopron	1 200		

8.5.12

Insertion of repeaters. No switched repeaters will be used.

8.6 TELEPHONE INFORMATION ABOUT HUNGARY

8.6.1 Toll Centres (See Table XV.)

8.6.2 Long-Distance Dialling Circuits (See Table XVI.)

TABLE XVI

Dialling Frequency Cycles per Second	Number of Circuits
50	53
2900	16
600/750	11 } Installed, but no longer in service
150	8

8.6.3 Subscriber Long-Distance Dialling

This will require seven digits; local service inside Budapest, six digits; and country districts, four digits. Digits "0" and "9" are reserved for long-distance service.

8.6.4 Call Duration

Local—2 minutes.
Toll—3–4 minutes.

8.6.5 Long-Distance Setting-Up Time

12–18 seconds.

8.6.6 Multi-Metering

It is planned that calls within an area will be multi-metered. Long-distance calls will be recorded automatically by ticket printers.

8.7 TELEPHONE INFORMATION ABOUT THE NETHERLANDS

8.7.1 National Numbering Scheme (See Table XVII.)

TABLE XVII

Toll Regional Centres	Code	Toll Terminal Areas
Goes	011	Middelburg, Zierikzee, Kruiningen, Oostburg, Ijzendijke, Terneuzen, Hulst
Breda	016	Tholen, Bergen op Zoom, Steenbergen, Roosendaal, Oudenbosch, Oosterhout.
's Gravenhage	017	Leiden, Alphen, Delft, Naaldwijk

TABLE XVII *Continued*

Toll Regional Centres	Code	Toll Terminal Areas
Rotterdam	018	Gouda, Gorinchem, Sliedrecht, Dordrecht, Oude Beijerland, Spijkenisse, Middelharnis
Alkmaar	022	Den Burg, Den Helder, Middenmeer, Schagen, Noord Scharwoude, Hoorn, Enkhuizen
Haarlem	025	Beverwijk, Zandvoort, Hillegom, Hoofddorp, Lisse
Amsterdam	029	Purmerend, Zaandam, Weesp, Amstelveem, Hilversum, Loenen
Utrecht	034	Harderwijk, Amersfoort, Tiel, Barneveld, Woerden, Culemborg, Leerdam, Jutphaas, Vreeswijk, Doorn
's Hertogenbosch	041	Tilburg, Waalwijk, Heusden, Zaltbommel, Oss, Veghel
Maastricht	044	Gulpen, Heerlen, Sittard
Venlo	047	Roermond, Helden-Panningen, Horst, Venray
Eindhoven	049	Eersel, Valkenswaard, Helmond, Deurne, Weert
Leeuwarden	051	Franeker, St. Anna Parochie, Dokkum, Veenwouden, Drachten, Sneek, Akkrum, Workum, Lemmer
Zwolle	052	Heerenveen, Wolvega, Steenwijk, Vollenhove, Meppel, Kampen, Ommen, Dedems Vaart, Elburg
Hengelo	054	Rijsen, Almelo, Goor, Neede, Oldenzaal, Groenlo, Enschede, Winterswijk
Beilen	055	Assen, Smilde, Borger, Emmen, Terapel, Hoogeveen, Oosterwolde, Coevorden
Groningen	059	De Leek, Winsum, Veendam, Middelstum, Appingedam, Hoogezand, Zuidhorn, Winschoten, Onstwedde
Deventer	067	Raalte, Epe, Apeldoorn, Zutphen, Vorden, Lochem, Uddel
Arnhem	083	Wageningen, Zevenaar, Dieren, Terborg, Doetinchem
Nijmegen	088	Cuijk, Grave, Druten, Zetten

8.7.2 Transmission

Although on some routes 2-wire lines are still in existence, 4-wire lines will be used throughout in the ultimate transmission plan.

Long-distance circuits are 12- or 24-channel carrier systems, which in future, and when necessary, may be extended to 36- or 48-channel systems. All these lines work on zero overall loss, and are used indiscriminately for terminating and tandem traffic, whenever the two types exist in one direction.

8.7.3 Routing

The latest plan foresees that, with a few exceptions, all district centres (total of 20 for the country) are directly interconnected by circuits with a high probability of loss. The overflow is alternatively routed through a common transit centre. The latter is directly connected to all district centres by circuits with a low probability of loss. The number of direct circuits is so calculated that some 5 or 10 per cent of the traffic overflow to a group of overflow circuits (common for all directions) which leads to one of the 20 district centres acting as an overflow tandem-exchange. This latter exchange is directly connected to all other district centres by a sufficient number of circuits (incoming and outgoing) to cater for the handling of all overflow traffic directed to it.

8.7.4 Numbering

The toll prefix comprises five figures, i.e., first figure "0" indicates toll call; combination of second and third figures indicates one of 20 districts into which the country is divided; the fourth figure indicates one of 10 sectors into which each district is divided; and the fifth figure indicates one of the 10 local exchange areas of each sector.

The local subscriber numbers may be 3, 4, 5, or 6 figures. The toll prefix is omitted for local calls exclusively.

8.7.5 Metering

The present metering is determined by the airline distances between the local-exchange areas and by time.

The fee may be determined completely by the prefix.

A new metering system is now under consideration, which is as follows. There will be only 3 zones, namely:

Zone *A*, comprising the home sector;

Zone *B*, applying to all adjacent sectors, in both the home and other districts; and

Zone *C*, comprising any other sectors. In this way the tariff may be determined by the first three figures of the prefix.

The numbering of metering impulses is as follows:

(*a*) At the moment of answering—1 impulse.

(*b*) After 5–10 seconds—a number of metering impulses which may be varied between 0 and 15, and which may be determined differently for each tariff, e.g., for Zone *A*—0 impulses, for Zone *B*—3 impulses, and for Zone *C*—10 impulses.

After this first period of 5–10 seconds has expired, one metering impulse is given for each time unit which is different in length for the 3 zones, e.g., for Zone *A*—1 impulse every 60 seconds, for Zone *B*—1 impulse every 30 seconds, and for Zone *C*—1 impulse every 10 seconds.

8.7.6 Time Limit

With the present tariff, the possibility is provided of breaking down the connection after 9 minutes; with the future tariff this breakdown may be done after a time varying between 6 and 12 minutes.

8.7.7 Traffic Metering

Each time- and zone-metering circuit is provided with two meters, which record, respectively, the total number of metering impulses and the total number of effective connections per circuit.

In order to obtain an indication of the ratio between the different kinds of calls, one time- and zone-metering circuit is provided with two totalling meters—one registering the total number of metering impulses for each tariff class, and the other meter registering the total number of effective calls per tariff class.

8.7.8 Register

Registers are at present provided only in rotary exchanges, but in future they will be introduced in all other exchanges and will take all figures dialled for toll calls and transmit them in tone-frequency code-impulse form.

8.7.9 Outlet Searching

In rotary exchanges all toll lines are in ideal groups; this condition is approached in step-by-step exchanges by the use of mixing selectors.

8.7.10 Selection Time

No limit fixed

8.7.11 Alternative Routing

See Section 8.7.3.

The possibility may be provided for routing calls through a second overflow exchange if all circuits to the first overflow exchange are occupied

8.7.12 Impulsing

Except on some 50-cycle-per-second junctions, all connections are set up via voice-frequency lines with end-to-end signalling, so that no impulse correctors are needed.

8.7.13 Signalling

On some existing circuits 50-cycle-per-second signalling is used, but on most circuits single voice-frequency signalling is applied, at 2400 cycles per second in one direction and 2500 cycles per second in the opposite direction.

8.7.14 Signals

First dialling tone (150 cycles per second).
Selection impulses.
Answering.
Forward clearing.
Release.
Offering:—

(*a*) Initiate offering.
(*b*) Acknowledgment of signal (*a*).
(*c*) Proceed offering } may be repeated.
(*d*) Undo offering }

Toll breakdown.
Re-ringing.
Clear back.
Second dialling tone after prefix (450 cycles per second).
Ringing control (450 cycles per second).
Busy tone (450 cycles per second).

8.7.15 Pulse Release

Incoming line—1–2 minutes.

8.7.16 Insertion of Repeaters

Terminal repeaters are used with pad switching.

8.7.17 4-Wire Switching

This is provided only on tandem connections.

8.7.18 Prefix for International Traffic

Not provided.

8.8 TELEPHONE INFORMATION ABOUT PORTUGAL

8.8.1 Transmission

Toll links are divided into separate groups for terminal and tandem traffic.

8.8.2 Tandem Routing

This has not yet been determined.

8.8.3 Metering

This will be determined by the code prefix.

8.8.4 Time Limit

Not yet decided.

8.8.5 Alternative Routing

This will probably be provided.

8.8.6 Prefix for International Traffic

Net yet determined.

8.8.7 Insertion of Repeaters

Probably terminating repeaters, and therefore no intermediate repeaters required.

8.8.8 Signalling

Trial installation with 600 and 750 cycles per second with compound prefix.

8.9 TELEPHONE INFORMATION ABOUT SWEDEN

8.9.1 Division of Country

300 areas which will require independent 3- or 4-figure codes commencing with the digit "0."

8.9.2 Local Numbering

Five or six figures, the first two figures indicating the identity of the exchange.

8.9.3 Open Numbering

Within each area only the local number is used; to reach other areas the code and local number must be dialled.

8.9.4 Tariff

The average price of a call is 4 öre. Experiments have been tried where the length of the call determines the fee. The basic length of conversation charge might be a period of 6 minutes.

8.9.5 Cordless Switchboard

Planned for Stockholm, and would probably be used elsewhere.

8.9.6 Toll Operator

Will set up long-distance connections with key set.

8.9.7 Priority

Is given to toll calls over local calls.

8.9.8 Repeaters

These will be automatically inserted.

8.9.9 4-Wire Switching

This will be adopted.

8.9.10 Transmission

The overall attenuation for a toll connection should not exceed 1.0 neper.

8.9.11 Toll Impulsing

Condenser discharge—50-cycle-per-second and 150-cycle-per-second alternating current; impulse—40 milliseconds, interruption—60 milliseconds.

8.9.12 National Numbering Scheme

An automatic national numbering scheme is planned.

8.10 TELEPHONE INFORMATION ABOUT SWITZERLAND

8.10.1 National Numbering Scheme (See Table XVIII.)

TABLE XVIII

Toll Regional Centres	Code	Toll Terminal Areas	Code
Lausanne	021	Genève (Primary)	022
		—	023
		Yverdon	024
		Aigle	025
		Martigny	026
		Sion	027
		Brig	028
		Bulle	029
Bern	031	Biel	032
		Thun	033
		B'dorf	034
		Langnau	035
		Interlaken	036
		Fribourg	037
		Neuchâtel	038
		La Chaux de Fonds	039
		Zweisimmen	030
Lucerne	041	Zug	042
		Schwyz	043
		Altdorf	044
		Sursee	045
Zürich	051	Winterthur	052
		Schaffhausen	053
		Frauenfeld	054
		Rapperswil	055
		Baden	056
		Wohlen	057
		Glarus	058
Basel	061	Olten (Primary)	062
		Langenthal	063
		Aarau	064
		S'thurn	065
		Delemont	066
St. Gallen	071	Weinfelden	072
		Wil	073
		Wattwil	074
Chur	081	St. Moritz	082
		Davos	083
		Schuis	084
		Sargano	085
		Ilanz	086
Lugano	091	Bellenzona	092
		Locarno	093
		Faido	094

8.10.2 Transmission

Primary, and in some cases secondary, toll links are divided into separate groups for terminal and tandem traffic.

8.10.3 Routing

If a secondary exchange is connected to two primary exchanges, both incoming and outgoing traffic has to be routed over the shorter route.

8.10.4 Tandem Routing

No connection should involve more than three primary toll exchanges.

8.10.5 Numbering

Four-, 5-, and 6-digit, and when necessary 3-digit, toll prefix.

8.10.6 Metering

Determined by distance between primary toll exchanges and therefore controlled by prefix. Facilities are included for determining fee by prefix and the two (or three) first figures of subscriber's number.

8.10.7 Time Limit

Facilities for applying forced release after 12 minutes on connection between primary toll centres; on secondary toll exchanges forced release may be applied after 6 minutes.

8.10.8 Traffic Metering

Totalling meter for day and night local connections.

Totalling meter for seizures per direction for outgoing terminal and tandem traffic in primary and secondary toll exchanges.

Overflow meters per direction in primary and secondary toll exchanges.

Totalling meters for seizures of originating outgoing primary and secondary toll traffic.

Totalling meters for transit connections in primary toll exchanges.

8.10.9 Registers

Arranged for 2 code and 6 numerical digits, code translation for 1–3 selections of 1–20 impulses and for controlled action with "seize" impulse and "end-of-selection" impulse. Impulses are transmitted in the forward direction.

8.10.10 Outlet Searching

Primary links are in ideal groups.

8.10.11 Selection Time

Maximum of 4 seconds.

8.10.12 Alternative Routing

Provided via alternative primary centre independent of register translation.

8.10.13 Impulsing

Primary toll centres are provided with impulse correctors.

8.10.14 Signalling

By 3000-cycle-per-second and 50-cycle-per-second alternating current.

8.10.15 Signals

Seize
Proceed-to-send.
Selection impulses.
End-of-selection impulses.
Answering.
Release.
Ringing control (400 cycles per second).
Busy tone (400 cycles per second).
Offering.
Re-ringing.
Clear-back.
Test busy.

8.10.16 Pulse Releases

Subscriber dialling	20–30 seconds
Incoming register	30 seconds
Incoming tandem line	5–10 seconds
Busy or ringing-tone condition	3 minutes

8.10.17 Insertion of Repeaters

By automatic means when connection exceeds 1.2 nepers.

8.10.18 4-Wire Switching

This is provided.

8.10.19 Prefix for International Traffic

014.

Shunted-Amplifier Input Admittance

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A METHOD is presented for finding the input admittance of an amplifier having a shunt impedance between its output and input. The shunted-amplifier gain is the sum of two terms that have simple physical significance. Several cases are worked out for a shunting capacitance, and include a single stage with both resistive and tuned loads and a double stage with resistance-capacitance coupling. Application to a frequency modulator is given in detail. Extensions are indicated to the theory of the tuned-plate tuned-grid oscillator, multivibrator, and negative-impedance devices.

• • •

When a passive network, such as a capacitance, is connected from input to output directly across a vacuum-tube amplifier, many of the amplifier characteristics may be radically altered. The input admittance, in particular, becomes a function of several variables; general and specific examples of its behavior have been discussed previously.[1–4]

The dependence of the input admittance of several closely related types of shunted amplifiers on input frequency, load impedance, and value of input-output shunting capacitance is examined. The ratio of the load impedance to the shunting impedance is introduced as the independent variable.

The circuit under study has been described[5] as a shunt-type feed-back amplifier. The action, so far as gain is concerned, is similar to that of the ordinary series-type feed-back circuit, in which a voltage from the output is applied in series with the input signal voltage. In the present case, if the generator is assumed to have zero internal impedance, the signal voltage from grid to cathode is specified by the assumed generator voltage. It cannot, therefore, be assumed to fluctuate with variation of the circuit parameters.

Two alternative assumptions are possible with respect to the effect of the shunt path:

A. Current which is dependent on the input voltage is introduced into the output.

B. Current which is dependent on the output voltage is introduced into the input.

The first assumption provides a logical basis for gain calculations; the second for calculation of input admittance.

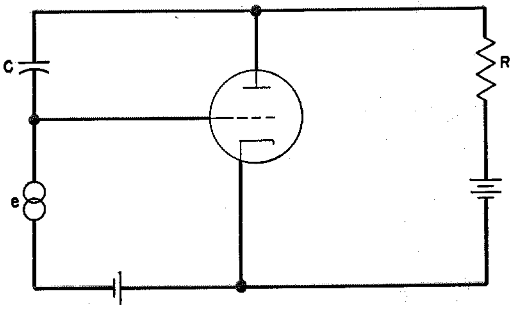

Fig. 1—Single-stage class-*A* amplifier with resistive plate load and with a capacitance *C* connected between input and output circuits.

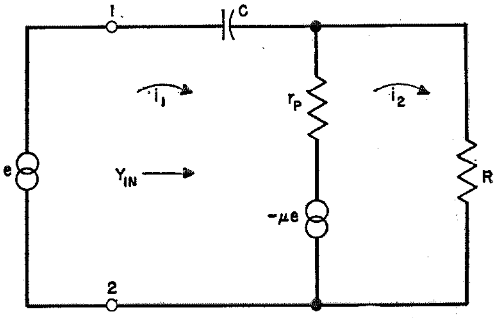

Fig. 2—Equivalent circuit of Fig. 1.

The theory developed below is applied to explain the operation of a particular frequency modulator[6] and is extended to the treatment of several circuits as negative-resistance oscillators.

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[1] J. M. Miller, "The Input Impedance of a Vacuum Tube," Bulletin No. 351, U. S. Bureau of Standards; 1919.

[2] S. Ballantine, "On the Input Impedance of the Thermionic Amplifier," *Physical Review*, v. 15, pp. 409–420; 1920.

[3] K. McIlwain and J. G. Brainerd, "High Frequency Alternating Currents," 2nd edition, John Wiley and Sons, New York, New York, 1939; p. 134.

[4] F. E. Terman, "Radio Engineering," 2nd edition, McGraw-Hill Book Company, New York, New York, 1937; p. 231.

[5] H. W. Bode, "Network Analysis and Feedback Amplifier Design," 1st edition, D. Van Nostrand and Company, New York, New York, 1945; p. 36.

[6] E. M. Ostlund, A. R. Vallarino, and M. Silver, "Center-Frequency-Stabilized Frequency-Modulation System," *Proceedings of the I.R.E.*, v. 35, pp. 1144–1149; October, 1947.

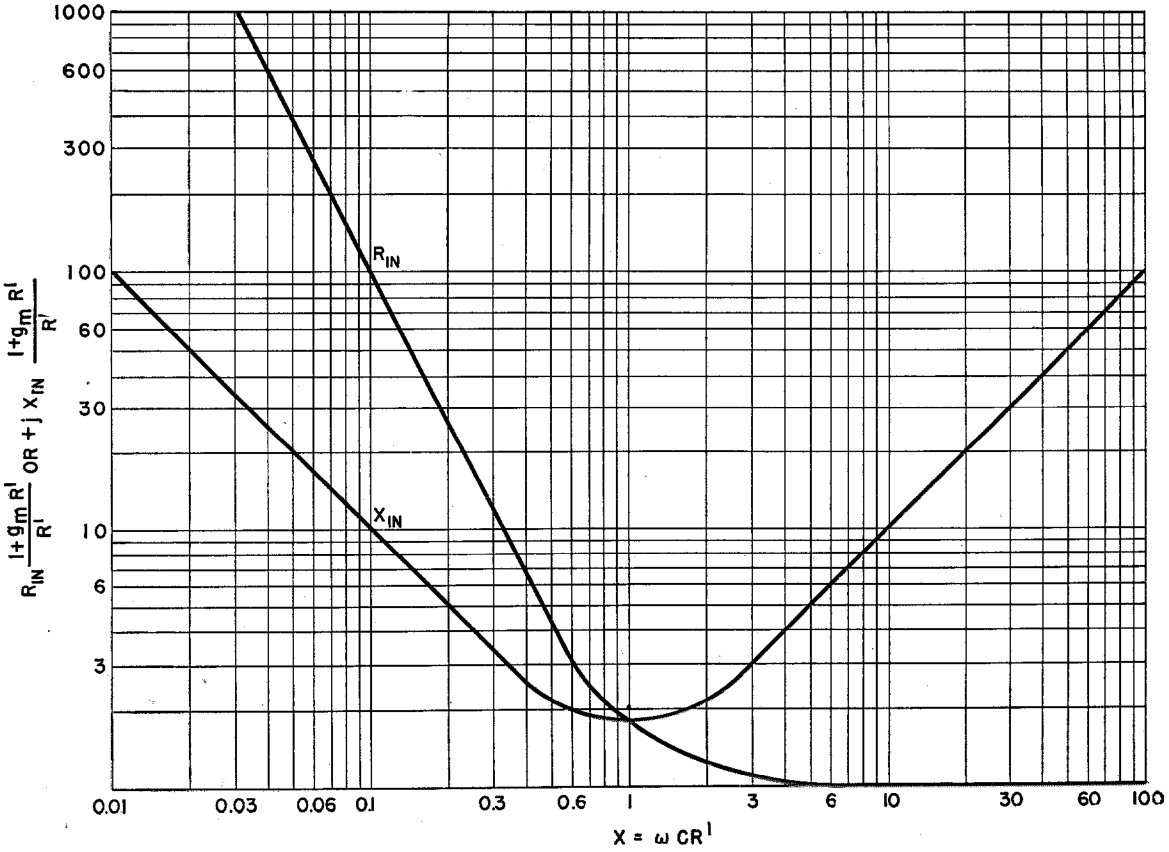

Fig. 3—Input resistance and reactance of circuit of Fig. 1.

1. *Single Stage With Resistive Load*

Fig. 1 is a schematic of a single-stage class-A amplifier with a capacitance connected between input and output, the amplifier being driven by a zero-impedance generator. The equivalent alternating-current circuit for frequencies below the transit-time domain is shown in Fig. 2. The grid-plate capacitance is included in C. The grid-cathode and plate-cathode capacitances may be neglected: the former constitutes a constant term additive to the input admittance, while the only effect of the latter is to limit (in practice but not in principle) the range of value of the load resistance. The input admittance Y_{in} between points 1 and 2 is

$$Y_{in}=\frac{i_1}{e}=\frac{r_p+R(1+\mu)}{r_pR+(r_p+R)\dfrac{-j}{\omega C}}. \tag{1}$$

Setting

$$R'=\frac{r_pR}{r_p+R},$$

$$x=\omega CR',$$

and

$$g_m=\frac{\mu}{r_p},$$

(1) may be written

$$Y_{in}=\frac{1+g_mR'}{R'}\times\frac{x^2+jx}{1+x^2}. \tag{2}$$

If Y_{in} is represented as a resistance R_{in} in parallel with a reactance X_{in}, then

$$R_{in}=\frac{R'}{1+g_mR'}\times\frac{1+x^2}{x^2}, \tag{3}$$

$$X_{in}=-j\frac{R'}{1+g_mR'}\times\frac{1+x^2}{x}. \tag{4}$$

Graphs of (3) and (4) are given in Fig. 3.

The gain A of the stage is given by Ri_2/e.

So,

$$A=\frac{\dfrac{Rr_p}{R+r_p}}{\dfrac{Rr_p}{R+r_p}+\dfrac{-j}{\omega C}}+\frac{-\mu\dfrac{R\frac{-j}{\omega C}}{R+\frac{-j}{\omega C}}}{r_p+\dfrac{R\frac{-j}{\omega C}}{R+\frac{-j}{\omega C}}}.$$

The gain is therefore the sum of two terms:

A. *Voltage division;* the parallel combination of load impedance and plate resistance, in series with the grid-plate impedance, divides the voltage impressed on this circuit.

B. *Tube amplification;* the tube load consists of the load impedance in parallel with the grid-plate impedance.

This is a general method of computing gain, and is used later in the paper. Thus,

$$A=\left(\frac{x^2+g_m^2R'^2}{x^2+1}\right)^{\frac{1}{2}}\tan^{-1}\left[\frac{(1+g_mR')x}{x^2-g_mR'}\right]. \quad (5)$$

It is evident that if

$$x\ll 1, \quad \text{then} \quad A\rightarrow -g_mR',$$

while when

$$x\gg 1, \quad A\rightarrow +1.$$

It should be noted that the above analysis is not strictly valid at frequencies considerably above approximately 20 megacycles per second for the following reasons:

A. The reactances of the various lead inductances become appreciable, so that Fig. 2 does not remain a true representation of Fig. 1. The main effect of these inductances is that the input admittance acquires a conductance component proportional to $\omega^2 g_m$.

B. Transit-time effects become important. Here also the input admittance acquires a conductance component proportional to $\omega^2 g_m$.

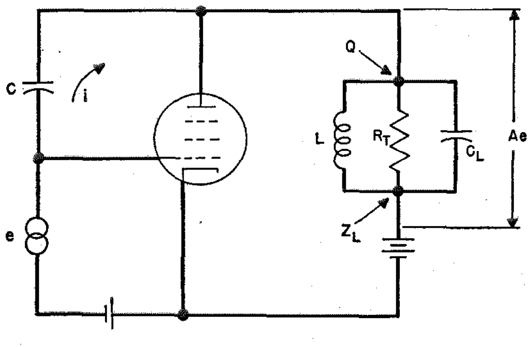

Fig. 4—Single-stage class-*A* amplifier with a tuned-circuit plate load and with a capacitance *C* connected between input and output circuits.

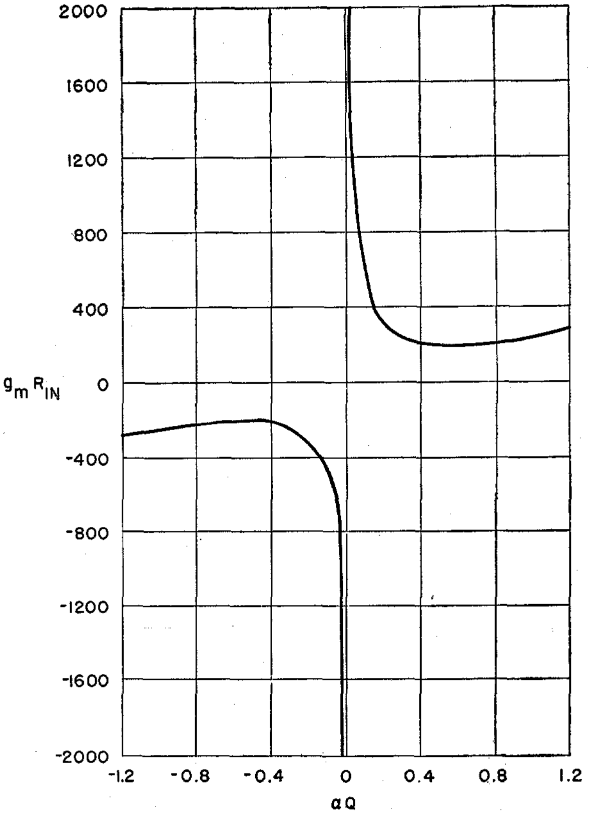

Fig. 5—Input resistance of circuit of Fig. 4 for $g_mR_T=20$ and $x=0.01$.

1.1 SMALL COUPLING

Consider first the case where C is very small, i.e.,

$$\frac{1}{\omega C}\gg\frac{r_pR}{r_p+R}, \quad \text{or} \quad x\ll 1.$$

The shunt input resistance R_{in} is then much larger than the shunt input reactance X_{in}, so the former may be neglected. It is interesting to note from (3), however, that when $g_mR'\gg 1$ and $x\ll 1$,

$$R_{in}=\frac{1}{g_m}\times\frac{1}{\omega^2C^2R'^2},$$

or that the input conductance varies as $\omega^2 g_m$. This is similar to the result of the lead-inductance and transit-time effects noted above. From (4), with $x\ll 1$,

$$X_{in}=-j\frac{R'}{1+g_mR'}\times\frac{1}{\omega CR'}$$

$$=\frac{-j}{\omega C(1+g_mR')}, \quad (6)$$

which is the reactance of a capacitor of value $C_{in}=C(1+g_mR')$. But it was shown above that if $x\ll 1$, then $A=-g_mR'$. Thus,

$$C_{in}=C(1-A). \tag{7}$$

This is the well-known formula for the Miller effect, which this case actually represents, the input admittance being a capacitance, the value of which depends on A. By placing the input terminals across the tank of an inductance-capacitance oscillator, or across a capacitance in a resistance-capacitance oscillator, the instantaneous effective capacitance may be varied, and the oscillator thereby becomes frequency modulated.

1.2 LARGE COUPLING

When C is very large,

$$\frac{1}{\omega C}\ll\frac{r_pR}{r_p+R}, \quad \text{or} \quad x\gg 1.$$

From Fig. 3 it is evident that the shunt input reactance X_{in} is now much larger than the shunt input resistance R_{in}, so the former may be neglected. Then, from (3),

$$R_{in}=\frac{R'}{1+g_mR'}. \tag{8}$$

If $g_mR'\gg 1$,

$$R_{in}=\frac{1}{g_m}. \tag{9}$$

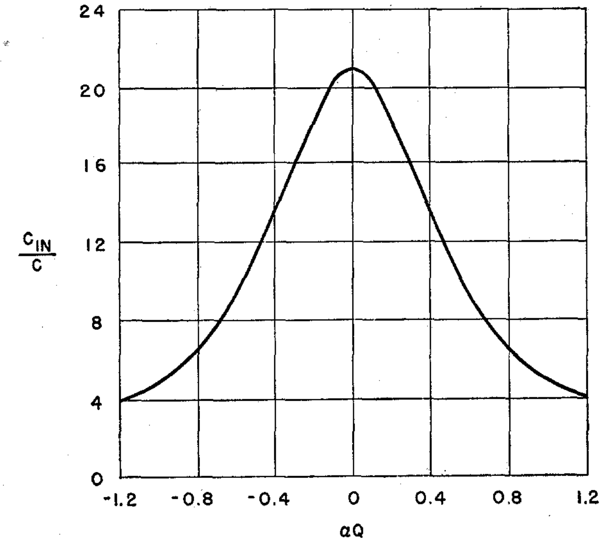

Fig. 6—Input capacitance of circuit of Fig. 4 for $g_mR_T=20$ and $x=0.01$.

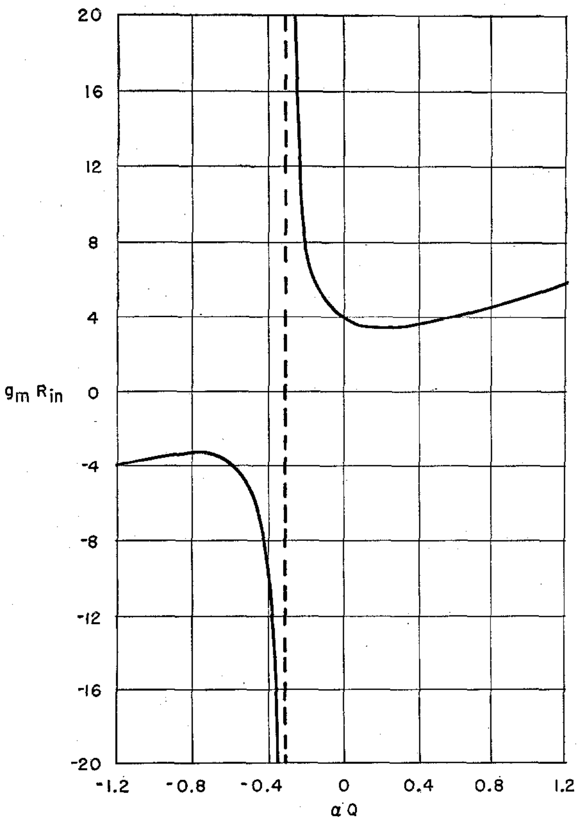

Fig. 7—Input resistance of circuit of Fig. 4 for $g_mR_T=20$ and $x=0.6$.

The input admittance in this case is a conductance, the value of which is proportional to g_m. This is for $x\gg 1$, which is the case not only when the coupling capacitor is made large, but also when the frequency is high. The behavior of this conductance with g_m is similar to that noted above for lead inductance and transit time; here, however, there is no longer a variation with frequency.

Since the input admittance is a resistance, the value of which depends on g_m, it is possible to vary the instantaneous frequency of a resistance-capacitance oscillator by placing the input terminals of the amplifier across the grid resistor of the oscillator. By using a pentode instead of a triode for the amplifier and varying g_m by applying audio-frequency voltage to the screen rather than to the control grid, the variable resistance may be obtained without accompanying voltage variations which could amplitude modulate the oscillator. This occurs because the amplifier gain, when $x\gg 1$, is unity regardless of the

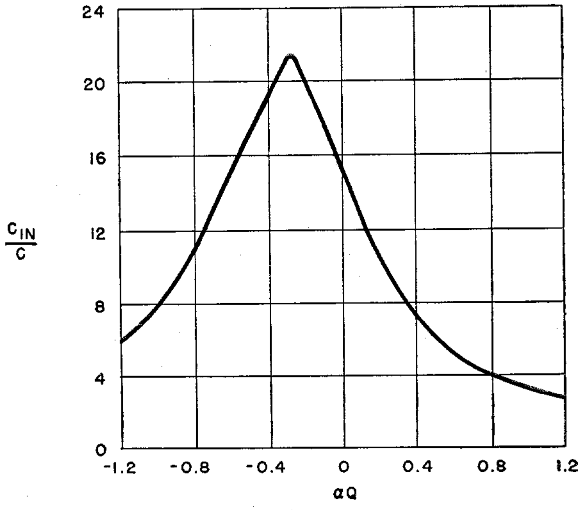

Fig. 8—Input capacitance of circuit of Fig. 4 for $g_mR_T=20$ and $x=0.6$.

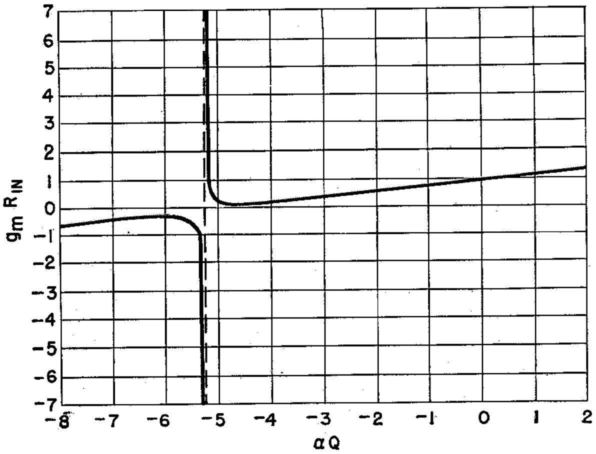

Fig. 9—Input resistance of circuit of Fig. 4 for $g_mR_T=20$ and $x=10$.

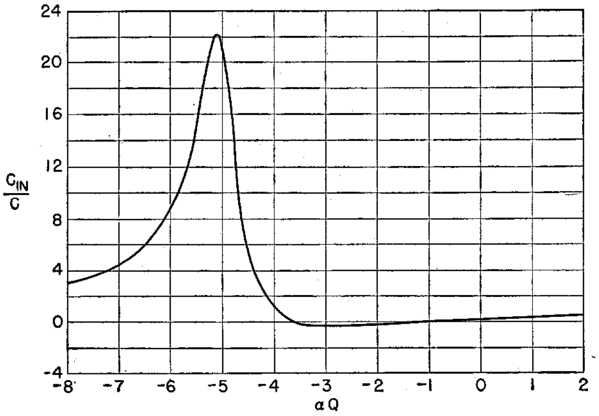

Fig. 10—Input capacitance of circuit of Fig. 4 for $g_mR_T=20$ and $x=10$.

value of g_m; no audio-frequency voltage can then appear at either grid or plate. To minimize amplitude modulation caused by the changing resistance itself, the input of the modulator may be placed across both resistors of the resistance-capacitance oscillator, rather than across the grid resistor alone.

2. *Single Stage Having Tuned-Circuit Load*

Fig. 1 may be modified by the substitution of a tuned circuit for the plate load resistor, as shown in Fig. 4. Fig. 3 is then no longer a valid representation of this new circuit, although it still holds true for any given frequency at which the tuned circuit is resonant.

From Fig. 4,

$$Y_{in}=\frac{i}{e}=\frac{1}{e}\left[\frac{e-Ae}{\frac{-j}{\omega C}}\right]=j\omega C(1-A). \qquad (10)$$

To calculate the gain A for an arbitrary C, the method derived above is used. If $\alpha=(\omega-\omega_0)/\omega_0$, where ω_0 represents the resonant angular velocity of the load alone, and discussion is restricted to the region near resonance, the load impedance is

$$Z_L=\frac{R_T}{1+j2\alpha Q}, \qquad (11)$$

where

$$Q=R_t/\omega_0 L.$$

If we set $x=\omega CR_T$ (assuming r_p large compared to R_T), we obtain

$$A=\frac{x^2+x2\alpha Q-g_mR_T+jx(g_mR_T+1)+2\alpha Qg_mR_T}{x^2+x4\alpha Q+4\alpha^2Q^2+1}. \qquad (12)$$

Substituting in (10),

$$R_{in}=\frac{x^2+x4\alpha Q+4\alpha^2Q^2+1}{x^2\left(g_m+\frac{1}{R_T}\right)+x2\alpha Qg_m}, \qquad (13)$$

and

$$C_{in}=C\frac{x2\alpha Q+4\alpha^2Q^2+1+g_mR_T}{x^2+x4\alpha Q+4\alpha^2Q^2+1}. \qquad (14)$$

These equations do not lend themselves to simple analysis as they stand. By limiting the range of applicability, however, it is possible to simplify them greatly.

2.1 $\alpha=0$

If the response at the resonant frequency of the tuned circuit is considered for various values of coupling, i.e., $\alpha=0$, the results are the same as those given in (3) and (4) for a resistive load. Fig. 3, therefore, applies in this case.

2.2 $x \ll 1$

Next is considered a region near the resonant frequency of the tuned circuit for small coupling between grid and plate, i.e., $x \ll 1$. This case is shown in Figs. 5 and 6, where $x=0.01$, $g_mR_T=20$. In order that ω shall not enter into both the abscissas and ordinates, these graphs imply that the plate tuning is changed from one point on the abscissa to another.

2.3 $x=0.6$

Next, let the coupling have some median value, such as $x=0.6$, and assume that $g_mR_T \gg 1$. This case is shown in Figs. 7 and 8 for the value $g_mR_T=20$.

2.4 $x \gg 1$

Finally, when the coupling is very large, and $g_mR_T \gg 1$, the case is shown in Figs. 9 and 10 for the values $x=10$ and $g_mR_T=20$.

3. *Frequency Modulator*

Fig. 11 is a schematic diagram of a frequency-modulated oscillator,[6] based on the principles just outlined. Several details of the modulator action may be of interest.

The frequency of operation, 4 megacycles, the grid-plate capacitance of 10 micromicrofarads, and the modulator load resistor of 2500 ohms are such that the value of x is approximately 0.6. Tuning of the modulator plate circuit is accomplished by inserting an audio-frequency voltage at the grid of the modulator. Correct tuning is then characterized by minimum audio-frequency voltage across the radio-frequency-bypassed part of the oscillator grid-leak resistor.

A pentode modulator with external capacitance is used, rather than a triode with its internal grid-plate capacitance, because of the greater linearity of the pentode g_m-e_g curve. The operating point is at the center of the most linear portion of the curve. The radio-frequency and audio-frequency voltages must never be so large that their sum extends beyond this linear region. For this reason, the modulator grid is tapped down on the oscillator coil (this connection is also used to minimize the effect of the modulator input-resistance variations).

Since operation is in the linear portion of the g_m-e_g curve, it is also in the curved portion of the i_p-e_g curve. The 6AB7 therefore acts as a van der Bijl grid modulator, and its plate-circuit

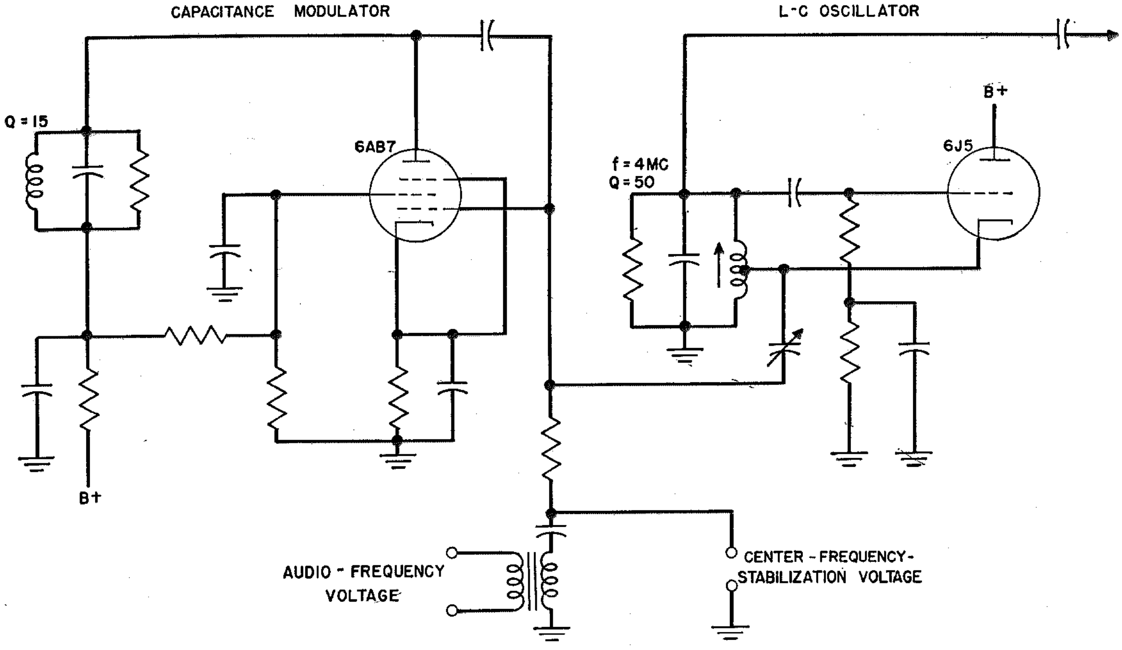
CAPACITANCE MODULATOR
L-C OSCILLATOR
Q = 15
6AB7
f = 4MC
Q = 50
B+
6J5
B+
AUDIO - FREQUENCY VOLTAGE
CENTER - FREQUENCY - STABILIZATION VOLTAGE

Fig. 11—Frequency-modulated oscillator.

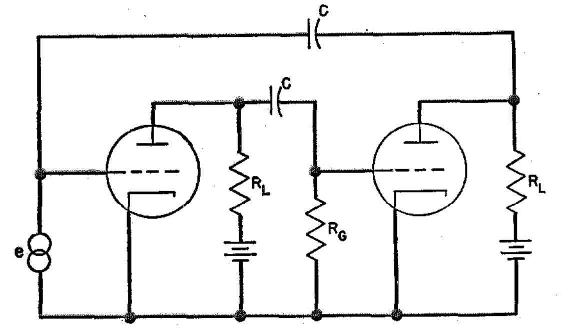

Fig. 12—Two-stage resistance-capacitance coupled class-*A* amplifier with a capacitance *C* connected between input and output circuits.

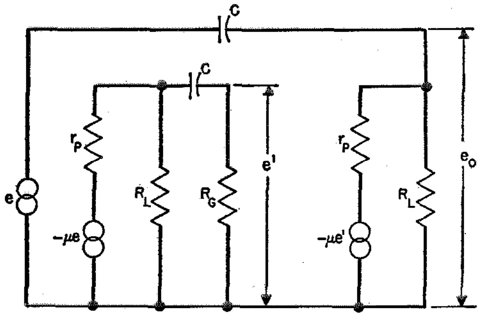

Fig. 13—Equivalent circuit of Fig. 12.

voltage will show the radio-frequency carrier plus the audio-frequency sidebands. The plate voltage is amplitude modulated as well as frequency modulated. Furthermore, amplitude components at twice the audio frequency are introduced into the plate voltage by the slight variation of the tuned-circuit impedance as the frequency varies on both sides of resonance. If, in addition, the plate-dropping, cathode, and screen bypass capacitors have a low reactance at radio frequency, but not at audio frequency, additional audio-frequency components are introduced into the plate voltage.

All these audio-frequency addition terms in the modulator plate voltage do not affect the voltage at the modulator grid since the grid-plate capacitor is very small. The audio-frequency modulated terms (radio-frequency sidebands), however, do feed back. The oscillator, therefore, is slightly amplitude modulated by the presence of the amplitude-modulated terms in the plate voltage of the modulator. Another way of saying the same thing is that both the modulator C_{in} and

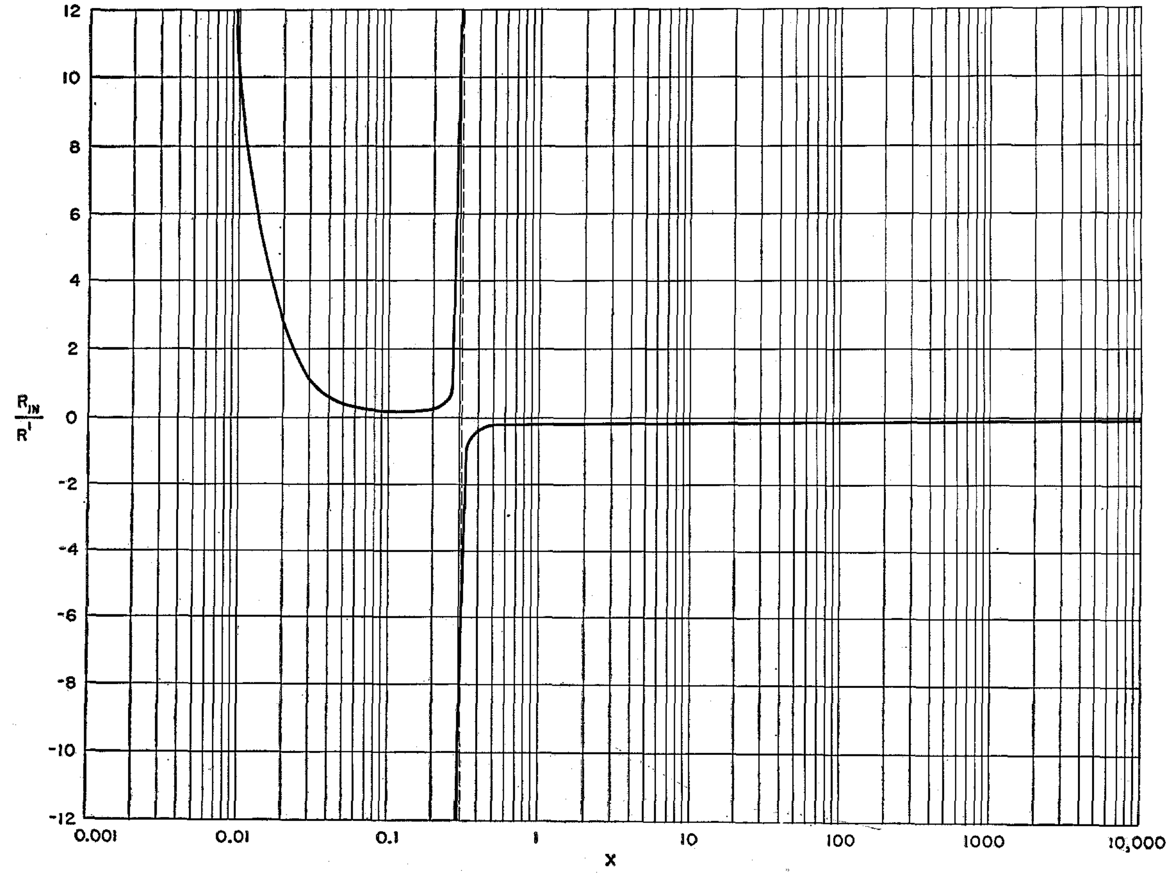

Fig. 14—Input resistance of circuit of Fig. 12 for $a=10$, $b=1000$, and $d=0$.

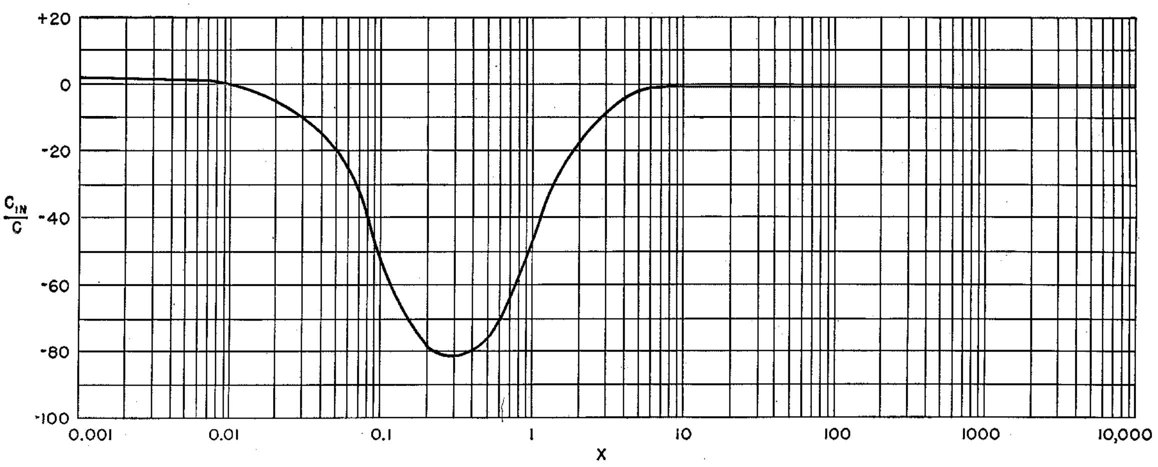

Fig. 15—Input capacitance of circuit of Fig. 12 for $a=10$, $b=1000$, and $d=0$.

R_{in} appear across the oscillator tuned circuit, the Q of which is varied with fluctuations of these terms, thus amplitude modulating the output. The effect is a small one, of the order of -50 decibels, if the modulator plate circuit has a low Q ($Q\approx 15$).

There are numerous ways in which the frequency deviation may be controlled: the audio-frequency level, the capacitance from 6AB7 grid to 6J5 cathode, the feed-back capacitance of the 6AB7, the Q of 6AB7 tuned circuit, etc.

The modulator distortion at 100-percent modulation, during a test of such a modulator, was approximately -46 decibels. The noise was -75 decibels.

4. *Tuned-Plate Tuned-Grid Oscillator*

The equations for input admittance developed in the previous sections may be applied for a rough explanation of the behavior of a tuned-plate tuned-grid oscillator, assuming that operation is class A rather than class C.

Equations (18) and (19), which give the input resistance and capacitance of the circuit of Fig. 4, may be written, if we let $\gamma=\omega/\omega_0$ and $K=\omega_0 CR_T$,

$$g_mR_{in}=1+\frac{Q}{K}-\frac{\frac{Q}{K}}{\gamma^2}+\frac{\frac{1+\frac{1}{g_mR_T}}{QK}}{\gamma^2\left(1+\frac{K}{Q}\right)-1}, \tag{15}$$

$$\frac{C_{in}}{C}=\frac{Q(\gamma^2-1)}{\gamma^2(K+Q)-Q+\frac{\gamma^2}{\gamma^2(K+Q)-Q}+\frac{\gamma^2g_mR_T}{[\gamma^2(K+Q)-Q]^2+\gamma^2}}. \tag{16}$$

If a parallel resonant circuit is now placed across the input of this circuit, then for stable oscillations to occur the following condition must be satisfied

$$R=-R_{in}. \tag{17}$$

From this condition, it is possible to show that three conditions may occur: no oscillation, one possible frequency of oscillation, and two possible oscillation frequencies. The values of these frequencies, as well as the value of inductance required in the grid, may also be calculated.

5. *Two-Stage Amplifier with Resistance-Capacitance Coupling*

A circuit of this type is shown in Fig. 12. The equivalent circuit, shown in Fig. 13, is based on assumptions of negligible capacitance between grid and plate of each stage, negligible capacitance from each grid and plate to ground, and class-A operation for each stage. Let the reactance of C be X_c. If

$$\left.\begin{aligned}R'&=\frac{R_Lr_p}{R_L+r_p},\\ x&=\omega CR',\\ a&=\frac{R_G}{R'},\\ b&=g_m^2R_GR',\end{aligned}\right\} \tag{18}$$

then Figs. 14 and 15 show the variation of input resistance and capacitance for the values $a=10$, $b=1000$, corresponding, e.g., to $R_G=10^5$, $R'=10^4$, and $g_m=10^{-3}$.

If it is desired to determine the frequency of stable oscillation when a resistance R_G is

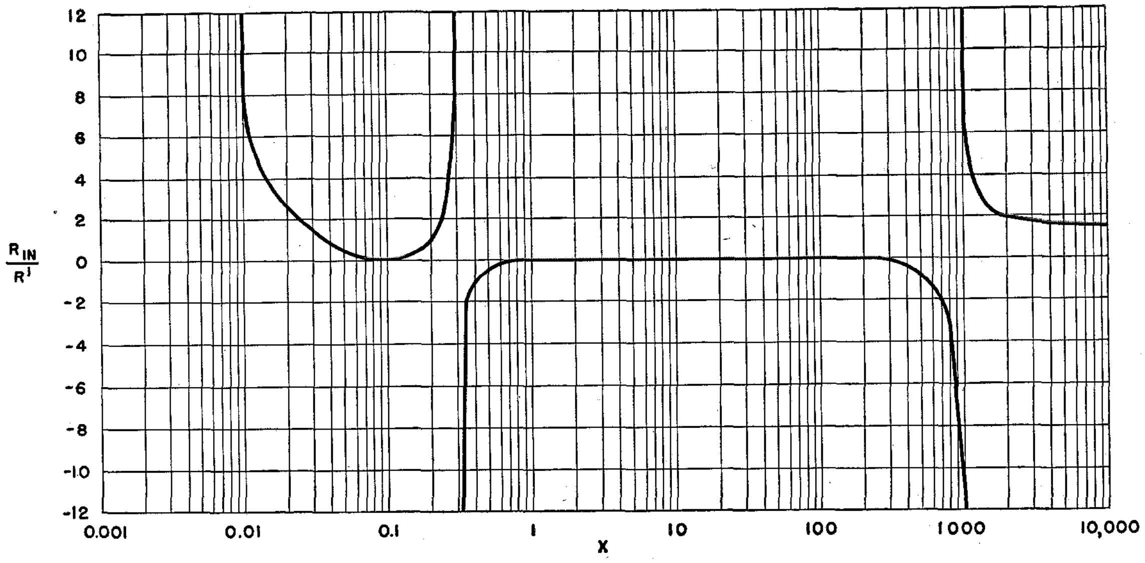

Fig. 16—Input resistance of circuit of Fig. 12 with capacitance C_1 between second grid and ground for $a = 10$, $b = 1000$, and $d = 0.01$.

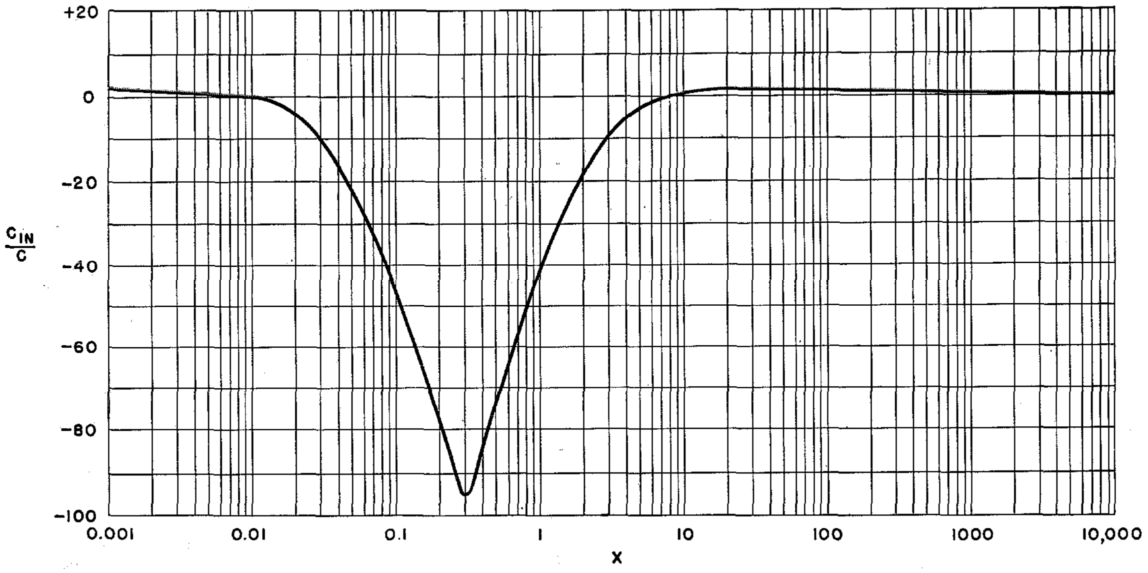

Fig. 17—Input capacitance of circuit of Fig. 12 with capacitance C_1 between second grid and ground for $a = 10$, $b = 1000$, and $d = 0.01$.

substituted for the generator in Fig. 12, the conditions to be met are

$$\frac{C_{in}}{C} = 0, \tag{19}$$

$$\frac{R_{in}}{R'} = -a. \tag{20}$$

These two conditions are incompatible in the present case: R_{in}/R' is positive when $C_{in}/C = 0$. Thus no oscillations will occur. It is shown below, however, that it is only necessary to make a simple assumption to permit oscillation.

If, for example, a capacitance C_1 of reactance X_1 is assumed to exist between the second grid and ground (in parallel with R_G in Fig. 12), and we let $d = X_c/X_1$, with R', x, a, and b having the same meaning as before, then the variation of input resistance and capacitance is shown in

Figs. 16 and 17 with $d=0.01$. It is seen that the conditions (19) and (20) for oscillation can now be fulfilled.

Previous analysis,[7] assuming a small inductance in series with one of the leads, has yielded a similar result. It was shown that if this inductance were zero, oscillations could not occur. Actually, it is immaterial whether series inductance or shunt capacitance is assumed; both are physically present.

6. Conclusion

The method outlined above for obtaining the input admittance of a shunted amplifier can be extended to many cases of practical importance; for example, oscillations in multistage intermediate-frequency amplifiers, the analysis of negative-resistance circuits such as are used to give a high variable Q in some wave analyzers, Eccles-Jordan center circuits, etc. The gain of the shunted amplifier is also obtained in simple fashion by this method.

In all these cases, by presupposing a generator of zero internal impedance between grid and cathode, it is possible to make the mathematical analysis without resorting to the concept of feedback.

7. Acknowledgment

The writer wishes to acknowledge the help of Messrs. E. M. Ostlund, M. Silver, and A. R. Vallarino on various phases of this work.

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[7] B. van der Pol, "Relaxation Oscillations," *Philosophical Magazine*, 7th series, v. 2, pp. 978–992; November, 1926.

Physical Properties of Crystals and Their Symmetry

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PREDICTION of certain possible properties of crystalline materials is based on the development of moduli linking the cause and effect of physical phenomena. The use of tensor and matrix calculus permits a survey of crystalline substances for such properties as pyroelectricity, dielectric susceptance, piezoelectricity, elasticity, and piezomagnetism. The physical properties of bodies crystallizing in various systems are outlined.

• • •

Pierre Curie, at the age of 21, published a short note[1] in which he announced the discovery of piezoelectricity.

The origin of this discovery must be sought in the associations of ideas caused by the simultaneous study of physics and crystallography.

"These two sciences were equally familiar to him and complemented each other in his mind.

"For him, the symmetry of phenomena was an intuitive notion. Few physicists had as great a knowledge as he of the crystallographic forms and of the symmetry groups."[2]

These main ideas, which led him to the discovery of the new phenomenon, were published later in the following form:

When certain causes produce certain effects, the symmetry elements of the causes must be found again in the effects.

When certain effects show a certain nonsymmetry, this nonsymmetry must be found again in the causes which produced them.

The reverse of these two propositions is not true, at least practically, i.e., the effects produced may be more symmetrical than their causes.

The discovery of piezoelectricity was brought about by the following reasoning. Pyroelectricity, a phenomenon that has been known for a long time, is characterized, in tourmaline, by the appearance, when heated, of a dielectric polarization.

This polarization can be represented by a vectorial quantity having the symmetry of a polar vector, although it is caused by a rise in temperature, which is a scalar phenomenon.

The nonsymmetry peculiar to the electric field (direction and space orientation) can appear only in a medium having this kind of nonsymmetry, i.e., with no center.

Consequently, other causes than a temperature variation, for instance, elastic deformations caused in the crystalline medium if the symmetry of this medium is favorable, will also be able to cause the appearance of a dielectric polarization.

Such was the origin of Curie's discovery, which was brilliantly confirmed by experiment. Curie found piezoelectric samples in the 21 classes of crystals having no symmetry centers. A detailed study of piezoelectricity was limited to quartz.

A few years later, Voigt, the German physicist who succeeded Neumann in this field, made a general investigation of analytical methods of representing the symmetry of the physical phenomena taking place in crystallized matter. These methods led Voigt to the notion of tensors. The very great physical and mathematical importance of tensorial calculus, this new mode of expressing the relation between quantities, is well known.

Voigt was able to foresee the existence of all piezoelectric moduli of crystals having no centers. He applied his analytical methods to all physical phenomena in crystals including elasticity, dielectric constants, piezomagnetism, etc. Present requirements and conditions suggest consideration of the use of crystals other than quartz. To know how a given piezoelectric crystal can be used, it is necessary to know the existence and the meaning of the characteristic moduli of its crystal class. These data are given very completely and in great detail by Voigt.

However, when trying to retrace the calculations on which Voigt based his conclusions, one is hampered by the notations he used, these notations being ill adapted to the problem.

Since 1910, when Voigt's work was done, mathematicians have made available to physicists greatly improved methods of calculation by means of which all the fundamental results of crystal physics can easily be found. For instance, by using appropriate notations, all theorems commonly applied in the study of symmetry can be obtained readily by the elementary rules of matrix multiplication. It is also possible to predict how a crystal, identified only by its main symmetry characteristics, must behave with respect to any physical property. Attention should first be directed to the theories that are indispensable to these calculations, which reduce to systematic use of linear forms and transformations.

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[1] Pierre Curie, "Développement, par pression, de l'électricité polaire dans les cristaux hémièdres à faces inclinées" (Development by Pressure of Polar Electricity in Hemihedral Crystals With Inclined Faces), *Comptes Rendus de l'Académie des Sciences;* August 2, 1880.

[2] Mme Curie in the preface to Pierre Curie's works.

The theory of tensors is obtained by associating the theories of linear transformations and of invariants. A tensor is a quantity characterized by the existence of several *numbers* (*components*) having particular values in a given reference system. However, any quantity defined by a system of components is not necessarily a tensor.

The main criterion of a tensor is the following:

With the components of a tensor, and in a manner which shall be defined more precisely later, it is possible to form an invariant.

If the reference system is changed, the tensor components also change, but the quantity represented by these components remains the same. It must be possible to calculate the new values of the components if the laws of transformation from the first system of reference to the second are known.

With regard to equality between tensors:

A. If two tensors are equal in one reference system, they are equal in all systems.

B. If a tensor is zero in one system, it is also zero in all others.

The relation between tensors is independent of the reference system. All the relations between physical quantities have this absolute characteristic of being independent of the reference system. This is commonly expressed by the notion of dimensions of physical quantities and of homogeneity of equations.

With crystalline media, the method of tensorial calculus becomes a necessity; the properties of symmetry will be characterized by the identity of the tensorial forms representing the quantities studied at two symmetrical points on the crystal.

1. Linear Transformations

Assume a group of variables of the same kind, a_1, a_2, a_3, ..., a_i they will be subjected to a linear transformation by calculating a new group of variables $\bar{a}_1$, $\bar{a}_2$, $\bar{a}_3$, ..., $\bar{a}_n$ by the following rule:

To obtain $\bar{a}_1$, for instance, a_1 is multiplied by a numerical coefficient α, then a_2 is multiplied by a second coefficient, and so on. The sum of all the products thus formed will equal $\bar{a}_1$.

The α coefficient figuring in the expression for $\bar{a}_1$ shall be provided with two indices, a lower index equal to 1 and an upper index equal to the index of the variable that this coefficient multiplies.

$$\bar{a}_1 = \alpha_1^1 a_1 + \alpha_1^2 a_2 + \dots + \alpha_1^n a_n. \tag{1}$$

The transformation will include n equations similar to (1). They may be written in an abridged form as

$$\bar{a}_1 = \sum_j \alpha_1^j a_j. \tag{2}$$

1.1 MUTE INDICES, EINSTEIN'S CONVENTION

If, in one member of an equation, the same index appears in the two terms of one product, as an upper index in one term and as a lower index in the other, we agree to effect the summation with respect to this index and it is unnecessary to retain the $\sum$ sign, and therefore, any value may be given to the index with respect to which the summation is effected. Such an index is called a mute index; (1) will be written

$$\bar{a}_i = \alpha_i^j a_j. \tag{3}$$

1.2 REVERSE TRANSFORMATION

To subject a new group of variables $\bar{a}_1$, $\bar{a}_2$, ..., $\bar{a}_n$, to the reverse linear transformation, coefficients β_m^n are selected to produce the initial variables a_1, a_2, ..., a_n.

This transformation is written:

$$a_m = \beta_m^n \bar{a}_n. \tag{4}$$

The β coefficient may be calculated from the α coefficient.

Let us consider the system of (3). Supposing a_j unknown and $\bar{a}_m$ known, we solve this system with respect to the a_j variables and, applying

Cramer's rule for relations of the type

$$a_1=\frac{\begin{vmatrix} \bar{a}_1 & \alpha_1^2 & \alpha_1^3 & \cdots & \alpha_1^n \\ \bar{a}_2 & \alpha_2^2 & \alpha_2^3 & \cdots & \alpha_2^n \\ \cdots & \cdots & \cdots & \cdots & \\ \bar{a}_n & \alpha_n^2 & \alpha_n^3 & \cdots & \alpha_n^n \end{vmatrix}}{\begin{vmatrix} \alpha_1^1 & \alpha_1^2 & \alpha_1^3 & \cdots & \alpha_1^n \\ \alpha_2^1 & \alpha_2^2 & \alpha_2^3 & \cdots & \alpha_2^n \\ \cdots & \cdots & \cdots & \cdots & \\ \alpha_n^1 & \alpha_n^2 & \alpha_n^3 & \cdots & \alpha_n^n \end{vmatrix}} \tag{5}$$

and developing,

$$a_1=\frac{|M_1^1|\bar{a}_1}{|\alpha|}-\frac{|M_1^2|\bar{a}_2}{|\alpha|}\cdots\pm\frac{|M_1^n|\bar{a}_n}{|\alpha|}. \tag{6}$$

By identifying with the first part of (4),

$$\beta_1^j=(-1)^{1+j}\frac{M_j^1}{|\alpha|} \tag{7}$$

and, in general,

$$\beta_i^j=(-1)^{i+j}\frac{|M|_j^i}{|\alpha|}, \tag{8}$$

$|M|_j^i$ being the minor obtained from the α determinant from which the jth line and the first column have been deleted.

Thus, the β coefficients are known when the α coefficients are known, provided, however, that (4) has a solution, that is, that the array of the α's be square and that determinant $|\alpha|$ be other than zero.

1.3 VARIANCE

Let us consider other variables $b_1, b_2, \ldots, b_n$ which, by virtue of their definitions, must undergo at the same time as variables a_i a linear transformation with the same coefficients α_i^j as variables a_j, i.e., such that

$$\bar{b}_i=\alpha_i^j b_j. \tag{9}$$

Variables b_j are said to be *covariant.*

Suppose, now, that there exist other variables c^i which, at the same time as the a_i's become $\bar{a}_j$'s, are transformed into $\bar{c}^k$ in accordance with the relation

$$\bar{c}^i=\beta_k^i c^k, \tag{10}$$

that is, by using for their direct transformation the coefficients of the reverse transformation of the covariant quantities. The variables c will be called *contravariant.*

These variables use α_i^j as a coefficient for their reverse transformation.

$$c^k=\alpha_l^k\bar{c}^l. \tag{11}$$

Contravariant variables will be distinguished from covariant variables by placing their indices at the top.

1.4 INVARIANT

Consider now the result of multiplying each variable a_j taken in the initial group by a quantity A and adding the terms.

$$S=A^j a_j. \tag{12}$$

A linear form is obtained.

Let us assume that when variables a_j are subjected to the initial linear transformation, the sum S remains constant.

$$S=\bar{A}^k\bar{a}_k. \tag{13}$$

This implies that the quantities A_j are functions of such a nature that

$$A^j a_j\equiv\bar{A}^k\bar{a}_k\equiv\bar{A}^k\alpha_k^j a_j. \tag{14}$$

This relation gives n equations

$$A^j=\alpha_k^j\bar{A}^k, \tag{15}$$

which implies that the quantities A^j are contravariant.

The quantities A are said to be the components of a tensor of the first order, once contravariant. Such quantities are familiar and are vectors.

1.5 NOTE

If, in (4), which defines the reverse transformation, the quantities $\bar{a}$ are expressed as functions derived from the fundamental linear transformation (3), the following relation is obtained

$$a_m\equiv\beta_m^n\alpha_n^p a_p, \tag{16}$$

which must be verified for all values of index p, the latter varying from 1 to n. Thus, n distinct equations are obtained between the α and the β coefficients.

If, for instance, $p=m$

$$a_m=\beta_m^n\alpha_n^m a_m \tag{17}$$

or

$$\beta_m^n\alpha_n^m=1. \tag{18}$$

If p differs from m, there will be $n-1$ equations.

$$0=\beta_m^n\alpha_n^p. \tag{19}$$

Equations (18) and (19) can be grouped into a single form

$$\beta^n_m\alpha^p_n=\delta^p_m, \tag{20}$$

where δ^p_m (Kronecker's index) has a value of 0 for $m\neq p$ and of 1 for $m=p$.

Equation (8) is the solution of equation (19).

1.6 Invariant Multilinear Forms

Suppose that there must be considered a p series of covariant variables

$$\begin{matrix} a_1, a_2, \ldots., a_n \\ b_1, b_2, \ldots., b_n \end{matrix}$$

and a q series of contravariant variables

$$\begin{matrix} e^1, e^2, \ldots., e^n \\ f^1, f^2, \ldots., f^n. \end{matrix}$$

We shall form the products $a_ib_j \ldots. e^kf^l$ in all possible manners giving indices i, j, k, and l, all possible values from 1 to n, and then multiply each product by a quantity A designated by the indices of the terms of the product arranged in opposite positions, adding all the terms obtained.

$$S=A^{ij}_{mn}a_ib_j\ldots.e^mf^n\ldots.. \tag{21}$$

If the quantities A are such that the sum S remains invariant when the linear transformation is effected, that is, if they are transformed into a quantity A such that

$$S=\bar{A}^{pq::::}_{rs::::}\bar{a}_p\bar{b}_q\ldots.\bar{e}^r\bar{f}^s\ldots., \tag{22}$$

these quantities A are said to be the components of a tensor of the $(p+q)$th order, p times contravariant and q times covariant.

If, in (22), the quantities $\bar{a}$, $\bar{b}$, ..., $\bar{e}$, $\bar{f}$ are replaced by their expressions in terms of a, b, c, ..., e, f, and if the coefficients of the same products are identified termwise, n relations are found, allowing the calculation of quantities $\bar{A}$ in terms of the A's.

$$\bar{A}^{pq::::}_{rs::::}=\beta^p_i\beta^q_j\alpha^m_r\alpha^n_s\ldots.A^{ij::::}_{mn::::}. \tag{23}$$

These equations follow:

$$\bar{A}^{pq}_{rs}=\beta^p_i\beta^q_j\alpha^m_r\alpha^n_s\ldots.A^{ij}_{mn}.$$

It will be seen that this condensed writing brings a considerable simplification, for (23) represents n^{p+q} equations each having n^{p+q} terms in their right-hand members.

Thus, in elasticity, use is made of tensors having 4 indices, representing in 3-dimensional space, $3^4=81$ components.

2. *Tensorial Algebra*

2.1 Addition

If two tensors have the same variance, they can be added termwise.

$$t^{ik}_m=r^{ik}_m+S^{ik}_m. \tag{24}$$

To demonstrate that the components t^{ik}_m have a tensorial nature, it will be sufficient to show that by collecting them, an invariant can be produced.

As r^{ik}_m and s^{ik}_m are the components of a tensor, the invariants are:

$$R=r^{ik}_ma_ib_kc^m, \tag{25}$$

$$S=s^{ik}_ma_ib_kc^m, \tag{26}$$

and

$$R+S=T=(r^{ik}_m+s^{ik}_m)a_ib_kc^m=t^{ik}_ma_ib_kc^m. \tag{27}$$

2.2 Exterior Multiplication

Consider two tensors whose components are A^r_{pq} and B^{lm}_k; each of the components A can be multiplied successively by all the B components, thus obtaining a tensor C whose components are

$$C^{rlm}_{pqk}=A^r_{pq}\times B^{lm}_k. \tag{28}$$

By a reasoning identical with the above, it could be shown that the components C^{rlm}_{pqk} have a tensorial character.

2.3 Contraction

By giving particular values to indices p, q and l, m, for instance, terms for which the lower index k of the B's is equal to the upper index p of the A's and form the sum, we may obtain components having only 4 indices.

$$D^{lm}_{pq}=\sum_r A^r_{pq}B^{lm}_r. \tag{29}$$

To show that these components D are tensorial in character, it is sufficient to verify that they are transformed in accordance with the law

$$\bar{D}^{zt}_{xy}=\alpha^p_x\alpha^q_y\beta^z_l\beta^t_mD^{lm}_{pq}. \tag{30}$$

As

$$A^u_{xy}=\alpha^p_x\alpha^q_y\beta^u_rA^r_{pq} \tag{31}$$

and

$$B^{zt}_s=\alpha^k_s\beta^z_l\beta^t_mB^{lm}_k, \tag{32}$$

the product will have new components, also tensorial in character.

$$C^{ztu}_{xys} = \alpha^p_x \alpha^q_y \beta^u_r \alpha^k_s \beta^z_l \beta^t_m A^r_{pq} B^{lm}_k. \tag{33}$$

If all the terms for which $r=k$ are added with respect to r,

$$C^{ztu}_{xys} = \sum_r \alpha^p_x \alpha^q_y (\beta^u_r \alpha^r_s) \beta^z_l \beta^t_m A^r_{pq} B^{lm}_r \tag{34}$$

or

$$\sum_r \beta^u_r \alpha^r_s = S^u_s,$$

Kronecker's index, which is equal to 1 when $u=s$ and to 0 when $u \neq s$, and as

$$\sum_r A^r_{pq} B^{lm}_r = D^{lm}_{pq},$$

there remains

$$\bar{D}^{zt}_{xy} = \bar{C}^{ztu}_{xyu} = \alpha^p_x \alpha^q_y \beta^z_l \beta^t_m D^{lm}_{pq}. \tag{35}$$

Again, in the present example, we can contract once more by selecting from all the terms D^{lm}_{pq} those for which $p=l$ and by adding with respect to l obtain the components of a tensor only once covariant and only once contravariant.

$$E^m_p = \sum_l D^{lm}_{pl}. \tag{36}$$

Finally, we can again add all the terms such as E^m_q in which $m=q$ with respect to m and obtain an invariant

$$J = \sum_m E^m_m. \tag{37}$$

Finally

$$J = A^r_{pq} B^{rq}_r. \tag{38}$$

3. *Matrixes*

The α^j_i coefficients of linear transformation (3) can be arranged in a table, the line number corresponding to the lower index and the column number to the upper index; this table is called the matrix of the linear transformation.

$$\left\| \begin{array}{ccccc} \alpha^1_1 & \alpha^2_1 & \alpha^3_1 & \cdots & \alpha^n_1 \\ \alpha^1_2 & \alpha^2_2 & \alpha^3_2 & \cdots & \alpha^n_2 \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ \alpha^1_n & \alpha^2_n & \alpha^3_n & \cdots & \alpha^n_n \end{array} \right\| \tag{39}$$

This table must be placed between double bars to avoid confusion with the determinant of the coefficients, whose typographical arrangement is the same.

The matrix is only a symbol representing a series of operations, which are represented in another way by (3). The matrix is represented in an abridged form by

$$\|\alpha\| \quad \text{or} \quad (\alpha). \tag{40}$$

In the linear transformation for obtaining p quantities $\bar{a}$ from n quantities a, p may differ from n, and the matrix is very often rectangular.

The *diagonal* terms are the α^i_i; they form the main diagonal of the table of the α^i_j.

The matrix is *symmetrical* when the terms that are symmetrical with respect to the main diagonal are equal. For instance,

$$\begin{array}{|c|c|c|} \hline a & d & e \\ \hline d & b & f \\ \hline e & f & c \\ \hline \end{array}\ . \tag{41}$$

It is antisymmetrical when the terms that are symmetrical with respect to the main diagonal are equal but of opposite signs; then $\alpha_{ij} = -\alpha_{ji}$ and, consequently, $\alpha_{ii} = 0$. For instance,

$$\begin{array}{|c|c|c|} \hline & d & e \\ \hline -d & & f \\ \hline -e & -f & \\ \hline \end{array}\ . \tag{42}$$

A diagonal matrix is a matrix in which all terms are zero except the diagonal terms.

$$\begin{array}{|c|c|c|} \hline a & & \\ \hline & b & \\ \hline & & c \\ \hline \end{array}\ . \tag{43}$$

A scalar matrix is a diagonal matrix in which all terms are equal.

$$\begin{array}{|c|c|c|} \hline a & & \\ \hline & a & \\ \hline & & a \\ \hline \end{array}\ . \tag{44}$$

The unit matrix is the scalar matrix in which $a=1$.

$$(1)=\begin{array}{|c|c|c|}\hline 1 & & \\ \hline & 1 & \\ \hline & & 1 \\ \hline \end{array}\,. \tag{45}$$

4. *Matrix Algebra*

4.1 EQUALITY

Two matrixes are equal when their components having the same indices are equal.

$$\alpha_j^i=\gamma_j^i$$

or

$$(\alpha)=(\gamma).$$

4.2 SUM

If we consider two linear transformations of the same nature on the same variables $a_1 \ldots . a_n$, let

$$\left.\begin{aligned} &\text{let} & b_p&=\alpha_p^q a_q \\ &\text{and} & C_p&=\gamma_p^q a_p, \end{aligned}\right\} \tag{46}$$

we can add the equalities having the same indices and place the corresponding terms as factors, obtaining

$$d_p=b_p+c_p=(\alpha_p^q+\gamma_p^q)a_p, \tag{47}$$

which corresponds to the linear transformation

$$d_p=S_p^q a_p.$$

The matrix S_p^q is the sum of the matrixes α_p^q and γ_p^q.

$$(S)=(\alpha)+(\gamma). \tag{48}$$

4.3 PRODUCT OF TWO MATRIXES

If, following a first linear transformation which acting on variables a transformed them into $\bar{a}$, we consider a second one transforming the $\bar{a}$'s into $\bar{\bar{a}}$'s,

let

$$\bar{a}_i=\alpha_i^j a_j, \tag{49}$$

$$\bar{\bar{a}}_k=\gamma_k^i \bar{a}_i, \tag{50}$$

we can note that the $\bar{a}_k$ terms are linear functions of the a variables.

$$\bar{\bar{a}}_k=\delta_k^j a_j. \tag{51}$$

Matrix (δ) is said to be the product of matrix (α) by matrix (γ).

$$(\delta)=(\alpha)(\gamma). \tag{52}$$

Replacing, in (50), the a_i's by their values from (49) and identifying termwise with (51), we see that

$$\delta_k^j=\sum_i \gamma_k^i \alpha_i^j. \tag{53}$$

This leads to the following rule for the multiplication of matrixes.

The two matrixes to be multiplied are placed side by side with the multiplier on the left of the multiplicand.

$$\overset{\longrightarrow}{\begin{array}{|c|c|c|}\hline \gamma_1^1 & \gamma_1^2 & \gamma_1^3 \\ \hline \gamma_2^1 & \gamma_2^2 & \gamma_2^3 \\ \hline \gamma_3^1 & \gamma_3^2 & \gamma_3^3 \\ \hline \end{array}} \;\times\; \downarrow \begin{array}{|c|c|c|}\hline \alpha_1^1 & \alpha_1^2 & \alpha_1^3 \\ \hline \alpha_2^1 & \alpha_2^2 & \alpha_2^3 \\ \hline \alpha_3^1 & \alpha_3^2 & \alpha_3^3 \\ \hline \end{array} = \begin{array}{|c|c|c|}\hline \gamma_1^1\alpha_1^1+\gamma_1^2\alpha_2^1+\gamma_1^3\alpha_3^1 & \gamma_1^1\alpha_1^2+\gamma_1^2\alpha_2^2+\gamma_1^3\alpha_3^2 & \\ \hline & & \\ \hline & & \\ \hline \end{array}\,. \tag{54}$$

To obtain the term δ_i^j, which is placed at the ith line and jth column of the product, we multiply successively each term of the ith line of the multiplier by the terms of the same number in the jth column of the multiplicand and add the products. The numbers are taken from left to right in the multiplier and from top to bottom in the multiplicand.

The result of this operation is not usually independent of the order of the factors. In general

$$(\alpha)(\gamma)\neq(\gamma)(\alpha). \tag{55}$$

4.3.1 Note 1

This practical rule applies to the linear transformation of (3). When it is desired to calculate the terms $\bar{a}$ as functions of the α's and of the a's, it is sufficient to arrange the a terms in one vertical column as shown in (56) and to operate in accordance with the rule for the multiplication of matrixes.

$$\begin{array}{|c|c|c|}\hline \alpha_1^1 & \alpha_1^2 & \alpha_1^3 \\ \hline \alpha_2^1 & \alpha_2^2 & \alpha_2^3 \\ \hline \alpha_3^1 & \alpha_3^2 & \alpha_3^3 \\ \hline \end{array} \times \begin{array}{|c|}\hline a_1 \\ \hline a_2 \\ \hline a_3 \\ \hline \end{array} = \begin{array}{|c|}\hline \alpha_1^1 a_1+\alpha_1^2 a_2+\alpha_1^3 a_3 \\ \hline \alpha_2^1 a_1+\alpha_2^2 a_2+\alpha_2^3 a_3 \\ \hline \alpha_3^1 a_1+\alpha_3^2 a_2+\alpha_3^3 a_3 \\ \hline \end{array} = \begin{array}{|c|}\hline \bar{a}_1 \\ \hline \bar{a}_2 \\ \hline \bar{a}_3 \\ \hline \end{array}. \qquad (56)$$

4.3.2 *Note 2*

The variables a_k, by means of which the basic transformation was defined, are of a tensorial character since by associating them with the contravariant variables b^k an invariant can be formed.

$$J=a_k b^k. \qquad (57)$$

The components a can be considered as being the components of a tensor of the first order, once covariant, i.e., a vector.

This tensor may be symbolized by an abridged symbol similar to that used for the matrix, and the basic transformation (3) will be written

$$(\bar{a})=(\alpha)(a). \qquad (58)$$

5. *Matrix Calculus Applied to Operations of Symmetry*

Operations of symmetry can easily be represented by linear transformation matrixes:

5.1 OPERATIONS OF THE FIRST KIND

In operations of the *first kind*, F passes to and can be made to coincide with $\bar{F}$. To bring about the coincidence of two geometrical figures, a translation and a rotation are combined.

In studying the symmetry of crystals, only rotation is considered. As the crystal is homogeneous, its properties depend only on orientation in space; directions alone need be considered.

The most general operation of the first kind is a rotation. If in Fig. 1 we pass from the coordinates of point M

$$M\left|\begin{array}{l} x^1 \\ x^2 \\ x^3 \end{array}\right.$$

to those of point $\bar{M}$, which is the new position of M when a rotation of angle θ is effected clockwise about axis ox^3, we find

$$\left|\begin{array}{l} \bar{x}^1=\rho\cos(\alpha-\theta) \\ \bar{x}^2=\rho\sin(\alpha-\theta) \\ \bar{x}^3=x^3 \end{array}\right. \qquad (59)$$

and, as $x^1=\rho\cos\alpha$, $x^2=\rho\sin\alpha$,

$$\left|\begin{array}{l} \bar{x}^1=x^1\cos\theta+x^2\sin\theta \\ \bar{x}^2=-x^1\sin\theta+x^2\cos\theta \\ \bar{x}^3=x^3. \end{array}\right. \qquad (60)$$

By analogy with the transformation relation of the contravariant quantities

$$\bar{x}^1=\beta_k^l x^k,$$

we see that transformation (60) can be represented by a matrix

$$B_k^i \rightarrow \begin{array}{|c|c|c|}\hline \cos\theta & \sin\theta & \\ \hline -\sin\theta & \cos\theta & \\ \hline & & 1 \\ \hline \end{array}. \qquad (61)$$

In what follows, we shall always consider the coordinates of a point as being contravariant

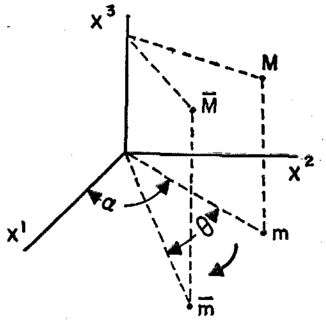


Fig. 1.

quantities. This follows the practice in vectorial geometry of using three unit vectors $\vec{a}_1$, $\vec{a}_2$, $\vec{a}_3$ to represent any vector that defines a reference system. The vector to be measured $\vec{F}$ is the resultant of three oriented vectors along the directions of the unit vectors, and each of these components is measured with the corresponding vector. Thus, we have an invariant sum,

$$\vec{F}=\vec{a}_i x^i.$$

The x^i's are called the coordinates of the extremity of vector θF and are contravariant. As the basic transformations consist in changing the unit vectors, the latter are naturally considered as being covariant.

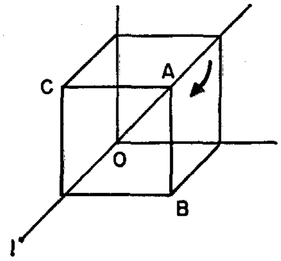


Fig. 2.

In the calculations which follow, we shall not use the methods of vectorial geometry, but those of metric geometry, the unit vectors common to all directions and to all systems shall be tacitly implied, but we shall continue to consider the coordinates of a point as contravariant quantities.

Furthermore, as we use rectangular axes, the matrix α of direct transformation of the covariant quantities is

$$\begin{array}{r|c|c|c|} \alpha_i^j\rightarrow & c & -s & \\ (\alpha)_3^\theta= & s & c & \\ & & & 1 \end{array}, \qquad (62)$$

$$(\alpha)_1^\theta=\begin{array}{|c|c|c|} 1 & & \\ & c & -s \\ & s & c \end{array}. \qquad (63)$$

It differs from matrix β only by a permutation of the lines and columns. For a simpler writing, we shall always write c and s instead of $\cos\theta$ and $\sin\theta$.

We shall designate the matrix by a lower index for the axis about which rotation takes place and an upper index indicating the angle of rotation.

A symmetry axis of order n is characterized by the value of $\theta=2\pi/n$.

If axis ox_3 is a binary axis, $\cos\pi=1$, $\sin\pi=0$,

$$(\alpha)_3^\pi=\begin{array}{|c|c|c|} -1 & & \\ & -1 & \\ & & 1 \end{array}, \qquad (64)$$

and

$$(\alpha)_1^\pi=\begin{array}{|c|c|c|} 1 & & \\ & -1 & \\ & & -1 \end{array}. \qquad (65)$$

In the study of crystals, we must obtain consider symmetry with respect to a ternary axis, Fig. 2, such as OA, oriented along the diagonal of a cube whose three sides coincide with the axes of coordinates.

The rotation, which changes point C to point B, corresponds to a circular permutation of the coordinates; the representative matrix is

$$(\alpha)_{\text{diagonal}}^{2\pi/3}=\begin{array}{|c|c|c|} & 1 & \\ & & 1 \\ 1 & & \end{array}. \qquad (66)$$

5.2 Operations of the Second Kind

In operations of symmetry of the *second kind*, F and $\bar{F}$ are in the same relation as an object and its mirror image. The figures are said to be enantiomorphous.

5.3 Symmetry With Respect to a Plane

We pass, in Fig. 3, from a point M to a point $\bar{M}$, symmetrical with respect to the normal plane

ox^3, by changing the sign of its coordinate x^3,

$$(\beta)_3^p=(\alpha)_3^p=\begin{array}{|c|c|c|}\hline 1 & & \\ \hline & 1 & \\ \hline & & -1 \\ \hline\end{array}\,. \tag{67}$$

Similarly, a symmetry with respect to the plane normal to direction 1 is represented by the matrix

$$(\beta)_1^p=(\alpha)_1^p=\begin{array}{|c|c|c|}\hline -1 & & \\ \hline & 1 & \\ \hline & & 1 \\ \hline\end{array}\,. \tag{68}$$

5.4 Symmetry With Respect to a Center

To pass from a point M to its symmetrical point $\bar{M}$ with respect to the center of coordinates, simply change the signs of its coordinates.

$$(\beta)_c=(\alpha)_c=\begin{array}{|c|c|c|}\hline -1 & & \\ \hline & -1 & \\ \hline & & -1 \\ \hline\end{array}=(-1), \tag{69}$$

i.e., the matrix (-1).

5.5 Combinations of Operations

Combinations of operations can be represented by a matrix equal to the product of the two operations. For example, if a point is rotated successively through an angle α and then through an angle β about an axis 3, the final operation is represented by the matrix corresponding to the rotation $(\alpha+\beta)$. This can easily be verified by the rule for the multiplication of matrixes.

$$(\alpha)_3^{\alpha+\beta}=(\alpha)_3^{\beta}\times(\alpha)_3^{\alpha}. \tag{70}$$

$$\begin{array}{|c|c|c|}\hline \cos\beta & -\sin\beta & \\ \hline \sin\beta & \cos\beta & \\ \hline & & 1 \\ \hline\end{array}\times\begin{array}{|c|c|c|}\hline \cos\alpha & -\sin\alpha & \\ \hline \sin\alpha & \cos\alpha & \\ \hline & & 1 \\ \hline\end{array}=\begin{array}{|c|c|c|}\hline \cos(\alpha+\beta) & -\sin(\alpha+\beta) & \\ \hline \sin(\alpha+\beta) & \cos(\alpha+\beta) & \\ \hline & & 1 \\ \hline\end{array}\,. \tag{71}$$

5.6 Plane of Alternate Symmetry of Order n

This is a plane direction, such that the crystal is restored to a position identical with its initial position by rotating it through π/n normal to the plane, and then taking its symmetrical point with respect to the plane. This involves two successive transformations, which are expressed by a matrix multiplication. For instance, for a plane of alternate symmetry of the second order,

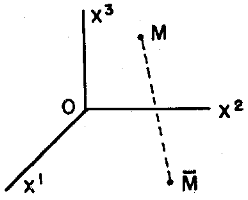


Fig. 3.

$$(\alpha)_3^{\bar{\omega}_2}=(\alpha)_2^p\times(\alpha)_3^{\pi/2}, \tag{72}$$

$$(\alpha)_3^{\bar{\omega}_2}=\begin{array}{|c|c|c|}\hline 1 & & \\ \hline & 1 & \\ \hline & & -1 \\ \hline\end{array}\times\begin{array}{|c|c|c|}\hline & -1 & \\ \hline 1 & & \\ \hline & & 1 \\ \hline\end{array}$$

$$=\begin{array}{|c|c|c|}\hline & -1 & \\ \hline 1 & & \\ \hline & & -1 \\ \hline\end{array}\,.$$

5.7 THEOREMS CONCERNING SYMMETRY

All theorems concerning symmetry may be proved by matrix multiplication.

5.7.1 *Theorem 1*

Any figure having an even-order axis L^{2n} passing through a symmetry center has a symmetry plane passing through the center and normal to the axis. Thus

$$(\alpha)^c \times (\alpha)_3^{\pi} = (\alpha)_3^{p}, \tag{73}$$

$$\underset{\text{Center}}{\begin{array}{|c|c|c|}\hline -1 & & \\ \hline & -1 & \\ \hline & & -1 \\ \hline \end{array}} \times \underset{\text{Axis}}{\begin{array}{|c|c|c|}\hline -1 & & \\ \hline & -1 & \\ \hline & & 1 \\ \hline \end{array}} = \underset{\text{Plane}}{\begin{array}{|c|c|c|}\hline 1 & & \\ \hline & 1 & \\ \hline & & -1 \\ \hline \end{array}}\,.$$

5.7.2 *Theorem 2*

Any figure having a plane of symmetry P passing through a center has an even-order axis L^{2n} normal to the plane and passing through the center.

$$\underset{\text{Center}}{\begin{array}{|c|c|c|}\hline -1 & & \\ \hline & -1 & \\ \hline & & 1 \\ \hline \end{array}} \times \underset{\text{Plane}}{\begin{array}{|c|c|c|}\hline 1 & & \\ \hline & 1 & \\ \hline & & -1 \\ \hline \end{array}} = \underset{\text{Axis}}{\begin{array}{|c|c|c|}\hline -1 & & \\ \hline & -1 & \\ \hline & & 1 \\ \hline \end{array}}\,. \tag{74}$$

5.7.3 *Theorem 3*

Any figure having a symmetry plane P and an even-order symmetry axis L^{2n} normal to the plane has a symmetry center P.

$$\underset{\text{Axis}}{\begin{array}{|c|c|c|}\hline -1 & & \\ \hline & -1 & \\ \hline & & 1 \\ \hline \end{array}} \times \underset{\text{Plane}}{\begin{array}{|c|c|c|}\hline 1 & & \\ \hline & 1 & \\ \hline & & -1 \\ \hline \end{array}} = \underset{\text{Center}}{\begin{array}{|c|c|c|}\hline -1 & & \\ \hline & -1 & \\ \hline & & -1 \\ \hline \end{array}}\,. \tag{75}$$

5.7.4 *Theorem 4*

The reverse occurs for odd-order axes.

$$\begin{array}{|c|c|c|}\hline \cos\dfrac{2\pi}{2n+1} & -\sin\dfrac{2\pi}{2n+1} & \\ \hline \sin\dfrac{2\pi}{2n+1} & \cos\dfrac{2\pi}{2n+1} & \\ \hline & & 1 \\ \hline \end{array} \times \begin{array}{|c|c|c|}\hline -1 & & \\ \hline & -1 & \\ \hline & & -1 \\ \hline \end{array} \neq \underset{\text{Plane}}{\begin{array}{|c|c|c|}\hline 1 & & \\ \hline & 1 & \\ \hline & & -1 \\ \hline \end{array}}\,. \tag{76}$$

5.7.5 Theorem 5

Any figure having an odd-order symmetry axis L_{2n+1} passing through a symmetry center C has a plane of alternate symmetry of order L_{2n+1}. As an example for $n=1$,

$$\begin{bmatrix} -1 & & \\ & -1 & \\ & & -1 \end{bmatrix} \times \begin{bmatrix} \cos\frac{2\pi}{3} & -\sin\frac{2\pi}{3} & \\ \sin\frac{2\pi}{3} & \cos\frac{2\pi}{3} & \\ & & \end{bmatrix} = \begin{bmatrix} \cos\frac{\pi}{3} & -\sin\frac{\pi}{3} & \\ \sin\frac{\pi}{3} & \cos\frac{\pi}{3} & \\ & & -1 \end{bmatrix} = (\alpha)^{\bar{\omega}_3}. \tag{77}$$

5.7.6 Theorem 6

Any figure having an alternate symmetry plane of odd order $\bar{\omega}_{2n+1}$ and, consequently, perpendicular axis of order $(2n+1)$ has a symmetry center.

$$(\alpha)^{\pi/(2n+1)} \times (\alpha)^{\bar{\omega}_3} = (-1). \tag{78}$$

$$\begin{bmatrix} \cos\frac{2\pi n}{2n+1} & -\sin\frac{2\pi n}{2n+1} & \\ \sin\frac{2\pi n}{2n+1} & \cos\frac{2\pi n}{2n+1} & \\ & & 1 \end{bmatrix} \times \begin{bmatrix} \cos\frac{\pi}{2n+1} & -\sin\frac{\pi}{2n+1} & \\ \sin\frac{\pi}{2n+1} & \cos\frac{\pi}{2n+1} & \\ & & -1 \end{bmatrix} = \begin{bmatrix} -1 & & \\ & -1 & \\ & & -1 \end{bmatrix}. \tag{79}$$

5.7.7 Theorem 7

Any figure having a plane of alternate symmetry of an even order $\bar{\omega}_{2n}$ and, consequently, an even axis L_{2n} can have no center.

$$\begin{bmatrix} \cos\frac{\pi}{2n} & -\sin\frac{\pi}{2n} & \\ \sin\frac{\pi}{2n} & \cos\frac{\pi}{2n} & \\ & & -1 \end{bmatrix} \times \begin{bmatrix} \cos\frac{\pi p}{2n} & \sin\frac{\pi p}{2n} & \\ \sin\frac{\pi p}{2n} & \cos\frac{\pi p}{2n} & \\ & & 1 \end{bmatrix} = \begin{bmatrix} \cos\frac{\pi(p+1)}{2n} & \sin\frac{\pi(p+1)}{2n} & \\ \sin\frac{\pi(p+1)}{2n} & \cos\frac{\pi(p+1)}{2n} & \\ & & -1 \end{bmatrix} \neq (-1), \tag{80}$$

as $\pi\frac{p+1}{2n} \neq \pi$.

6. *Symmetry in Crystals*

At two symmetrical points of a crystal, all physical properties are identical. By physical symmetry is meant a condition whereby if at point M certain causes produce certain physical effects, at a point $\bar{M}$, symmetrical with respect to the first point, the same causes produce the same effects. These relations between causes and effects are characterized by moduli, whose numerical values depend on the material.

In the analytical expression between causes and effects, space coordinates are involved. We are attempting to specify, therefore, in tensorial form, the dependence between causes, effects, and position coordinates of a point. As a result, the moduli are themselves tensors and when all moduli are known at point M, their new values at a different point $\bar{M}$ may be calculated.

If M and $\bar{M}$ are symmetrical and if the calculated moduli must be identical to the corresponding moduli as measured at M, as many relations will be obtained as there are moduli.

The whole group of relations thus established permit the assertion that certain moduli can or cannot exist within the limits of symmetry characteristic of the material studied. In some cases, it is found that certain moduli are equal to certain others with or without a change in signs.

In all cases, the equations between moduli are homogeneous equations of the type $ax+by+\ldots.=0$; no constant in the right-hand member ever indicates the order of magnitude of these moduli. This method of calculation allows only a prediction of the possible existence of the moduli and their mutual relations.

To predict the properties of all crystalline systems, it is necessary to examine only a few particular relations of symmetry; the combination of relations found permits the inherent characteristics of each crystalline species to be determined.

7. *Pyroelectricity, Tensorial Moduli With One Index*

This phenomenon corresponds in a dielectric medium to the appearance of an electrostatic-induction vector when the temperature of the body is varied.

In the general case of a dielectric placed in an electric field, the induction b has the value

$$\vec{b}=Ko\vec{h}+4\pi\vec{P}. \tag{81}$$

We can assume, here, that the external electric field is zero, and therefore

$$\vec{b}=4\pi\vec{P}. \tag{82}$$

Vector $\vec{P}$ is homogeneous to vector $\vec{b}$ and will have the same variance.

The work of an electric charge e moved over a distance $d\vec{x}$ in a medium in which the electric field is $\vec{h}$ is

$$dw=e\vec{h}d\vec{x}, \tag{83}$$

or, using the tensorial notation,

$$dw=eh_i dx^i,$$

where dw, as well as e, are invariants. We must thus consider h as a tensor of the first order, once covariant. The same holds for b in rectangular coordinates. The energy stored in volume V of a dielectric medium is

$$W=\frac{\vec{b}\vec{H}}{8\pi}dv,$$

$$8\pi W=b_i h_i dx^i d\sigma^i.$$

As b is proportional to the temperature variation θ in a pyroelectric body, this property must be characterized by moduli π that must satisfy

$$b_i=\pi_i x\theta. \tag{84}$$

As the two members of this equation must be tensors of the same order, it is evident that as θ is a scalar, i.e., a tensor of zero order, π must be a tensor of the first order, covariant.

Then, if we pass from one point M to another point $\bar{M}$, the moduli $\bar{\pi}$ at point $\bar{M}$ can be calculated in terms of the moduli π at point M by means of the linear relation

$$\bar{\pi}_k=\alpha_k^l\pi_l \tag{85}$$

and, moreover, if point $\bar{M}$ is symmetrical with respect to M, all physical properties at $\bar{M}$ are identical to those measured at M and, therefore,

$$\bar{\pi}_k=\pi_k. \tag{86}$$

By associating (85) and (86), we finally obtain the relation of the particular symmetry represented by the α coefficients,

$$\pi_k=\alpha_k^l\pi_l. \tag{87}$$

The linear transformation can be made very rapidly by matrix multiplication and (87) becomes

$$(\pi)=(\alpha)(\pi). \tag{88}$$

If, for instance, the system has a center of symmetry, the equation is written

$$\begin{bmatrix} \pi_1 \\ \pi_2 \\ \pi_3 \end{bmatrix} = \begin{bmatrix} -1 & & \\ & -1 & \\ & & -1 \end{bmatrix} \times \begin{bmatrix} \pi_1 \\ \pi_2 \\ \pi_3 \end{bmatrix} = \begin{bmatrix} -\pi_1 \\ -\pi_2 \\ -\pi_3 \end{bmatrix}. \tag{89}$$

Therefore:

$$\pi_1=-\pi_1=\pi_2=-\pi_2=\pi_3=-\pi_3=0. \qquad (90)$$

Pyroelectricity cannot exist in centered systems; all the moduli are zero.

The relations for symmetry about a nonbinary axis ox_3 will be written

$$(\pi)=(\alpha)_3^{\theta}(\pi) \qquad (91)$$

or

$$\begin{bmatrix}\pi_1\\ \pi_2\\ \pi_3\end{bmatrix}=\begin{bmatrix}c & -s & \\ s & c & \\ & & 1\end{bmatrix}\times\begin{bmatrix}\pi_1\\ \pi_2\\ \pi_3\end{bmatrix}=\begin{bmatrix}c\pi_1-s\pi_2\\ s\pi_1+c\pi_2\\ \pi_3\end{bmatrix}, \qquad (92)$$

which corresponds to equations

$$\left.\begin{aligned}\pi_1(c-1)&=s\pi_2\\ \pi_1 s&=-(c-1)\pi_2\\ \pi_3&=\pi_3\\ \text{or}\qquad (\pi_1)^2&=-(\pi_2)^2\\ \pi_3&=\pi_3.\end{aligned}\right\} \qquad (93)$$

The two first equations are verified only if moduli π_1 and π_2 are zero; the third one implies that π_3 *can* exist. The existence of modulus π_3 is not sure, but it is not in contradiction with the type of symmetry considered.

For symmetry about axis L_3, a diagonal of the cube, we have

$$\begin{bmatrix}\pi_1\\ \pi_2\\ \pi_3\end{bmatrix}=\begin{bmatrix} & 1 & \\ & & 1\\ 1 & & \end{bmatrix}\times\begin{bmatrix}\pi_1\\ \pi_2\\ \pi_3\end{bmatrix}=\begin{bmatrix}\pi_2\\ \pi_3\\ \pi_1\end{bmatrix}$$

or

$$\pi_1=\pi_2=\pi_3, \qquad (94)$$

three equal possible moduli of the same absolute value. In a crystal having only this symmetry, a rise in temperature would give three induction components, equal in the three directions ox_1, ox_2, ox_3, i.e., a resultant precisely in the direction of the diagonal of the cube.

7.1 ALTERNATE PLANE $\bar{\omega}_2$, PERPENDICULAR TO AXIS ox_3

The relations are:

$$\begin{bmatrix}\pi_1\\ \pi_2\\ \pi_3\end{bmatrix}=\begin{bmatrix} & -1 & \\ 1 & & \\ & & -1\end{bmatrix}\times\begin{bmatrix}\pi_1\\ \pi_2\\ \pi_3\end{bmatrix}=\begin{bmatrix}-\pi_2\\ \pi_1\\ \pi_3\end{bmatrix}. \qquad (95)$$

Hence, $\pi_1=\pi^2=-\pi^2=0$. π_3 alone can exist.

7.2 BINARY AXES OR PLANES OF SYMMETRY

If there exist still other axes or planes, these symmetries can reduce the number of moduli. The general relation for symmetry being

$$\pi_i=\alpha_i^k\pi_k, \qquad (96)$$

where the symmetrical elements are planes or axes, the transformation matrix is always diagonal, involves only coefficients of the type α_i^i, and is equal to $+1$ or to -1. Relation (96) can thus be written

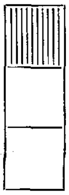
$$\pi_i=\pm\pi_i. \qquad (97)$$

Where the coefficient α_i^i is negative, the corresponding modulus cannot exist; where it is positive, however, pyroelectricity can appear in the corresponding direction.

Rectangular tables can be prepared in which + or − signs represent the result of matrix multiplication on a unit vector. The squares containing + signs may be cut out and the grid thus obtained superposed over tables of moduli. For example, the binary ox_1-axis case follows.

$$\begin{bmatrix}+1\\ +1\\ +1\end{bmatrix}\neq\begin{bmatrix}+1 & & \\ & -1 & \\ & & -1\end{bmatrix}\times\begin{bmatrix}+1\\ +1\\ +1\end{bmatrix}=\begin{bmatrix}+1\\ -1\\ -1\end{bmatrix}. \qquad (98)$$

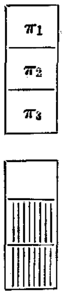
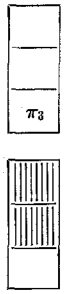
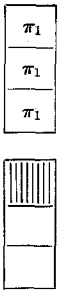
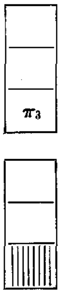
The top box containing +1 is cut out to produce the grid below.



The ox_1 axis already being transformed by the symmetries of the nonbinary axes ox_3 or, as the case may be, by the symmetries about the ternary axis or diagonal of the cube, the possible moduli for all crystalline systems will appear in the holes of the grid as they are the only ones consistent with all the symmetries of the system. (See Tables I and II.)

TABLE I

Grids for Use with Moduli Tables

No Symmetry	Nonbinary ox_3 Axis	Cube Diagonal	Alternate Plane
			
Plane Perpendicular to ox^1	Plane Perpendicular to ox^3	Binary Axis to ox	Binary Axis to ox^3

From the results for all crystalline systems, which will be given later, it is evident that pyroelectricity appears only in a few systems.

8. *Dielectric Susceptance, Tensorial Moduli With Two Indices*

If an isotropic dielectric is placed inside an electric field, the electric induction is equal to

$$\vec{b}=K_0\vec{h}+4\pi\vec{P}=K_1\vec{h}=K_0\vec{h}+4\pi x\vec{h}, \qquad (99)$$

where K_1 is the specific inductive power and x the dielectric susceptance.

In an anisotropic dielectric medium, the direction of the induction is not necessarily that of the field; the above relation will be written:

$$b_i=K_0h_i+4\pi x_i^k h_k, \qquad (100)$$

$$\frac{b_i-K_0h_i}{4\pi}=x_i^k h_k. \qquad (101)$$

The quantities $(b_i-K_0h_i)/4\pi$ and h_k are once covariant; hence all the x_i^k are components of a tensor of the same order, once covariant and once contravariant.

The tensorial transformation for calculating the susceptances x at a point $\bar{M}$, when they are known at point M, are

$$\bar{x}_i^j=\alpha_i^k\beta_l^j x_k^l, \qquad (102)$$

TABLE II

Moduli of 10 Crystalline Forms

System	Form	Moduli	Name	Symbol
Triclinic	Hemihedrism	π_1, π_2, π_3	Potassium Dichromate	$K_2Cr_2O_7$
Clinorhombic	Hemihedrism Antihemidrism	π_3 π_1, π_2		
Orthorhombic	Antihemidrism	π_3	Lithium Sulfate	Li_2SO_4
Rhombohedric	Antihemidrism Tetartohedrism	π_3	Scalclesite	—
Quadratic	Antihemidrism Tetartohedrism	π_3	Wulfenite	$PbMoO_4$
Hexagonal	Antihemidrism Tetartohedrism	π_3	Tourmaline Potassium Bromate Sodium Periodate	— $KBrO_3$ $NaIO_4$

and, if point $\bar{M}$ is symmetrical with respect to M, we must have the identities

$$\bar{x}_i^j = x_i^j, \tag{103}$$

hence the relation between the coefficients involved in a type of symmetry characterized by a particular matrix α is

$$\bar{x}_i^j = \alpha_i^k \beta_l^j x_k^l \tag{104}$$

and, as we are using rectangular axes and $\alpha_j^i = \beta_i^j$, these equations can be written

$$x_i^j = \sum_{kl} \alpha_i^k \alpha_j^l x_k^l. \tag{105}$$

It can easily be shown that a transformation of the type

$$a_{ij} = \alpha_i^k \alpha_j^l a_k^l,$$

to which (105) is similar, can be effected by matrix multiplication and

$$(a) = (\alpha) \times (a)(\alpha)^T. \tag{106}$$

Matrix $(\alpha)^T$ represents matrix α transposed, i.e., the lines and columns have been interchanged.

From the definition of a matrix product, obtaining the product $(\alpha)(a)$ involves finding terms such that

$$I_i^l = \alpha_i^k a_{kl}. \tag{107}$$

Then, obtaining the product $(I)(\alpha)^T$ involves finding terms such that

$$a_{ij} = I_i^l \alpha_l^j = \alpha_l^j \alpha_i^k a_k^l. \tag{108}$$

But α_l^j of the transposed matrix is equal to the term of the direct matrix. Transformation (105) can then be obtained by matrix multiplication (106),

$$(x) = (\alpha)(x)(\alpha)^T. \tag{109}$$

As in the case of one-index tensors, we have first found the relation imposed by symmetry about axis ox_3 for an angle different from π. We must next solve the system of equations.

$$\begin{array}{c|c|c|c|}
 & 1 & 2 & 3 \\ \cline{2-4}
1 & x_1^1 & x_1^2 & x_1^3 \\ \cline{2-4}
2 & x_2^1 & x_2^2 & x_2^3 \\ \cline{2-4}
3 & x_3^1 & x_3^2 & x_3^3 \\ \cline{2-4}
\end{array}
\equiv
\begin{array}{|c|c|c|}\hline
c & -s & \\ \hline
s & c & \\ \hline
 & & 1 \\ \hline
\end{array}
\times
\begin{array}{|c|c|c|}\hline
x_1^1 & x_1^2 & x_1^3 \\ \hline
x_2^1 & x_2^2 & x_2^3 \\ \hline
x_3^1 & x_3^2 & x_3^3 \\ \hline
\end{array}
\times
\begin{array}{|c|c|c|}\hline
c & s & \\ \hline
-s & c & \\ \hline
 & & 1 \\ \hline
\end{array}, \tag{110}$$

$$\begin{array}{|c|c|c|}\hline
c & -s & \\ \hline
s & c & \\ \hline
 & & 1 \\ \hline
\end{array}
\times
\begin{array}{|c|c|c|}\hline
x_1^1 & x_1^2 & x_1^3 \\ \hline
x_2^1 & x_2^2 & x_2^3 \\ \hline
x_3^1 & x_3^2 & x_3^3 \\ \hline
\end{array}
=
\begin{array}{|c|c|c|}\hline
x_1^1c - x_2^1s & x_1^2c - x_2^2s & x_1^3c - x_2^3s \\ \hline
x_1^1s + x_2^1c & x_1^2s + x_2^2c & x_1^3s + x_2^3c \\ \hline
x_3^1 & x_3^2 & x_3^3 \\ \hline
\end{array},$$

and

$$\begin{array}{|c|c|c|}\hline
x_1^1c - x_2^1s & x_1^2c - x_2^2s & x_1^3c - x_2^3s \\ \hline
x_1^1s + x_2^1c & x_1^2s + x_2^2c & x_1^3s + x_2^3c \\ \hline
x_3^1 & x_3^2 & x_3^3 \\ \hline
\end{array}
\times
\begin{array}{|c|c|c|}\hline
c & s & \\ \hline
-s & c & \\ \hline
 & & 1 \\ \hline
\end{array}
=
\begin{array}{|c|c|c|}\hline
\begin{array}{c} x_1^1c^2 - x_2^1cs \\ + x_2^2s^2 - x_1^2cs \end{array} & \begin{array}{c} x_1^1cs - x_2^1s^2 \\ - x_2^2cs + x_1^2c^2 \end{array} & x_1^3c - x_2^3s \\ \hline
\begin{array}{c} x_1^1cs + x_2^1c^2 \\ - x_2^2cs - x_1^2s^2 \end{array} & \begin{array}{c} x_1^1s^2 + x_2^1cs \\ + x_2^2c^2 + x_1^2cs \end{array} & x_1^3s + x_2^3c \\ \hline
x_3^1c - x_3^2s & x_3^1s + x_3^2c & x_3^3 \\ \hline
\end{array}.$$

Equating the terms of the initial tensor and of the right-hand member, we have the equations

$$\left.\begin{array}{ll}
(11) & (x_2^2-x_1^1)s=(x_1^2+x_2^1)c \\
(12) & -(x_2^2-x_1^1)c=(x_1^2+x_2^1)s \\
(13) & x_1^3(c-1)=x_2^3s \\
(23) & x_1^3s=-x_2^3(c-1) \\
(31) & x_3^1(c-1)=x_3^2s \\
(32) & x_3^1s=-x_3^2(c-1),
\end{array}\right\} \quad (111)$$

whence $x_1^1=x_2^2$, $x_1^2=-x_2^1$, $x_1^3=x_2^3=x_3^1=x_3^2=0$, and $x_3^3=x_3^3$. The symmetry conditions in the case of a two-index tensor for a nonbinary axis ox^3 correspond to the tensor

$$\begin{array}{|c|c|c|}
\hline
x_1^1 & x_1^2 & \\
\hline
-x_1^2 & x_1^1 & \\
\hline
 & & x_3^3 \\
\hline
\end{array}\;.$$

There are now only three distinct susceptances instead of nine.

8.1 SYMMETRY ABOUT THE TERNARY AXIS, DIAGONAL TO THE CUBE

The tensorial transformation

$$x_i^j=\sum_{kl}\alpha_i^k\alpha_j^l x_k^l$$

reduces to an interchange of indices in the right-hand member, for in the matrix

$$(\alpha)_{\text{diagonal}}^{2\pi/3}=\begin{array}{|c|c|c|}
\hline
 & 1 & \\
\hline
 & & 1 \\
\hline
1 & & \\
\hline
\end{array}\;, \qquad (66)$$

all the coefficients are equal to unity but the numbers of their lines are greater by 1 than the numbers of their columns. The symmetry conditions are

$$\begin{gathered}
x_i^j=x_{i+1}^{j+1} \\
x_1^1=x_2^2=x_3^3 \\
x_1^2=x_2^3=x_3^1 \\
x_1^3=x_2^1=x_3^2.
\end{gathered}$$

Then this tensor must be

$$\begin{array}{|c|c|c|}
\hline
x_1^1 & x_1^2 & x_1^3 \\
\hline
x_1^3 & x_1^1 & x_1^2 \\
\hline
x_1^2 & x_1^3 & x_1^1 \\
\hline
\end{array}\;.$$

8.2 ALTERNATE PLANE

We must write the equations

$$\begin{array}{|c|c|c|}
\hline
x_1^1 & x_1^2 & x_1^3 \\
\hline
x_2^1 & x_2^2 & x_2^3 \\
\hline
x_3^1 & x_3^2 & x_3^3 \\
\hline
\end{array}
\equiv
\begin{array}{|c|c|c|}
\hline
 & -1 & \\
\hline
1 & & \\
\hline
 & & -1 \\
\hline
\end{array}
\times
\begin{array}{|c|c|c|}
\hline
x_1^1 & x_1^2 & x_1^3 \\
\hline
x_2^1 & x_2^2 & x_2^3 \\
\hline
x_3^1 & x_3^2 & x_3^3 \\
\hline
\end{array}
\times
\begin{array}{|c|c|c|}
\hline
 & 1 & \\
\hline
-1 & & \\
\hline
 & & -1 \\
\hline
\end{array}\;.$$

The tensor readily reduces to

$$\begin{array}{|c|c|c|}
\hline
x_1^1 & x_1^2 & \\
\hline
-x_1^2 & x_1^1 & \\
\hline
 & & x_3^3 \\
\hline
\end{array}\;.$$

8.3 SYMMETRY ABOUT BINARY AXES AND WITH RESPECT TO PLANES PERPENDICULAR TO THE AXES OF COORDINATES

The reasoning is the same as in the case of tensors having two indices. Grids will be prepared in the same way with the boxes cut away (shaded in the text) for the terms +1 which correspond to possible moduli. The boxes are not shaded for factors −1, thus indicating that the corresponding susceptances are impossible. By superposing one or several of such grids on the four fundamental tables, it is possible to show which crystalline systems have susceptances that are not zero.

TABLES

No Symmetry

x_1^1	x_1^2	x_1^3
x_2^1	x_2^2	x_2^3
x_3^1	x_3^2	x_3^3

Nonbinary ox_3 Axis or Alternate Plane

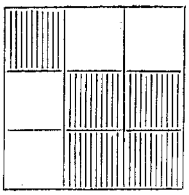
x_1^1	x_1^2	
$-x_1^2$	x_1^1	
		x_3^3

Cube Diagonal

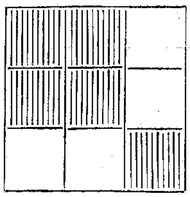
x_1^1	x_1^2	x_1^3
x_1^3	x_1^1	x_1^2
x_1^2	x_1^3	x_1^1

GRIDS

Axis L_1^2 and Plane P_1



Axis L_3^2 and Plane P_3



9. *Piezoelectricity, Tensorial Moduli With Three Indices*

Piezoelectricity is a phenomenon of interdependence between the elastic properties of a crystalline medium and electric induction in this medium.

This phenomenon has two inverse aspects. The first one, discovered by Curie, is that if a crystalline medium having no symmetry center is subjected to stresses, a dielectric polarization appears in the medium. The second one, which Lippmann had predicted from thermodynamic considerations as a consequence of the first, is the deformation of the medium when subjected to an electric field.

Before defining the various moduli possible in a material whose symmetry characteristics are known, we must define what we mean by deformations and by stresses in a solid body, and bring out the tensorial character of these quantities.

We shall consider first a plane deformation. In Fig. 4, let M and P be two neighboring points in a plane having the coordinates

$$M\left|\begin{array}{l} x^1 \\ x^2 \end{array}\right. \qquad P\left|\begin{array}{l} x^1+dx^1 \\ x^2+dx^2 \end{array}\right. \tag{112}$$

in the undeformed matrix. If the matrix is deformed, points M and P are displaced and move to M', P' having the coordinates

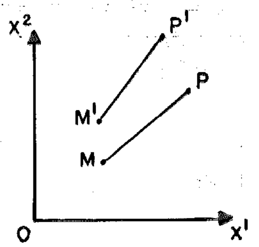


Fig. 4.

$$\left.\begin{array}{ll} M' & \left|\begin{array}{l} x^1+u^1 \\ x^2+u^2 \end{array}\right. \\ P' & \left|\begin{array}{l} x^1+u^1+dx^1+\dfrac{\partial u^1}{\partial x^1}dx^1+\dfrac{\partial u^1}{\partial x^2}dx^2, \\ x^2+u^2+dx^2+\dfrac{\partial u^2}{\partial x^1}dx^1+\dfrac{\partial u^2}{\partial x^2}dx^2. \end{array}\right. \end{array}\right\} \tag{113}$$

We see that vector MP has changed length as well as direction. The lengths $M'P'$ and MP are independent of the reference system as well as of any function depending on these lengths and, in particular, of the quantity:

$$\frac{\overline{M'P'}^2-\overline{MP}^2}{2}=J.$$

As a solid is not easily deformed, u^1 and u^2 are infinitely small. Also, the differential quotients $\partial u^1/\partial x^1$, $\partial u^1/\partial x^2$, $\partial u^2/\partial x^1$, $\partial u^2/\partial x^2$ are infinitely small quantities of the first order and, therefore, their squares or their products can be neglected. Consequently, we have

$$J=\frac{\partial u^1}{\partial x^1}(dx^1)^2+\frac{\partial u^2}{\partial x^2}(dx^2)^2+\left(\frac{\partial u^1}{\partial x^2}+\frac{\partial u^2}{\partial x^1}\right)dx^1dx^2. \tag{114}$$

$$S_{ij}=\frac{1}{2}\left(\frac{\partial u^i}{\partial x^j}+\frac{\partial u^j}{\partial x^i}\right)=S_{ji}. \tag{115}$$

The invariant J then appears as

$$J = S_{ij}\, dx^i\, dx^j. \tag{116}$$

We see that when $i=j=1$, for instance, $S_{11} = \frac{\partial u^1}{\partial x^1}$. S_{11} represents the relative lengthening in the direction 1. It can easily be verified by considering two small vectors of lengths du^1 and du^2 parallel to the axes before deformation, that the cosine of their angle after deformation is equal to

$$\cos\theta = \frac{\partial u^1}{\partial x^2} + \frac{\partial u^2}{\partial x^1}, \tag{117}$$

therefore,

$$S_{ij} = \tfrac{1}{2} \cos\theta_{ij}.$$

If points M and P had been taken outside the plane x^1ox^2, we should still have (116) but, in the summation we should have to make i, j successively equal to 1, 2, 3.

From the point of view of tensorial transformations, dx^1 and dx^j must be considered as contravariant variables, the nine quantities S_{ij} (of which only six are distinct since $S_{ij}=S_{ji}$) are the components of a second-order tensor twice covariant, and, knowing S_{ij} in one system, we can calculate S_{ij} for any system by means of the transformation relation

$$\bar{S}_{kl} = \alpha_k^i \alpha_l^j S_{ij}. \tag{118}$$

We can now conceive of the most general piezoelectric phenomenon as being the appearance of a deformation S_{ij} in a crystalline body placed in a field with a component h_k.

If this deformation occurs, assuming it to be linear, the coefficient of proportionality between S_{ij} and h_k may be designated by δ_{ij}^k; it is a piezoelectric modulus and we must have

$$S_{ij} = \delta_{ij}^k h_k. \tag{119}$$

Relation (119) has two very distinct meanings: First, if the values of i, j, and k are fixed, modulus δ_{ij}^k may be defined by a simultaneous measurement of S_{ij} and h_k. It should be noted that as

$$S_{ij} = S_{ji} \quad \text{and} \quad \delta_{ij}^k = \delta_{ji}^k,$$

there are really only 18 distinct moduli. However, for further discussion and because of the symmetry of notations, it is much more convenient to assume that there are nominally 27 moduli.

Second, if we consider k as a mute index, (119) represents 9 equations, each having three terms in its right-hand member.

As homogeneity is essential from the tensorial point of view, and as (119) represents a contracted multiplication, the 27 moduli S_{ij}^k are the components of a third-order tensor, twice covariant, and once contravariant.

This makes it possible, when the 27 moduli are known in one system, to calculate them in another system through the relation

$$\bar{\delta}_{mn}^p = \alpha_m^i \alpha_n^j \beta_k^p \delta_{ij}^k, \tag{120}$$

which in rectangular coordinates is

$$\bar{\delta}_{mn}^p = \sum_{ijk} \alpha_m^i \alpha_n^j \alpha_p^k \delta_{ij}^k. \tag{121}$$

These relations also represent the numerical values of moduli at a point $\bar{M}$ when they are known at point M.

In a medium completely without symmetry, as in the hemihedrism form of the triclinic system, all 27 moduli can exist. This means that an electric field oriented in any manner can cause the appearance of all the elastic deformations.

In a medium having symmetry, the necessary relations exist between the moduli so that, if $\bar{M}$ is symmetrical with respect to M, there must be 27 such relations.

$$\bar{\delta}_{mn}^p = \delta_{mn}^p, \tag{122}$$

and, by associating them with (120), the relations between moduli at point M are obtained.

$$\delta_{mn}^p = \sum_{ijk} \alpha_m^i \alpha_n^j \alpha_p^k \delta_{ij}^k, \tag{123}$$

the α's being the coefficients of the transformation matrix of the coordinates of one point in coordinates of the symmetrical point for the symmetry considered.

If the body has a center of symmetry, the transformation matrix is (-1), i.e.,

$$\alpha_1^1 = \alpha_2^2 = \alpha_3^3 = -1,$$

and (123) can be written

$$\delta^p_{mn} = \alpha^m_m \alpha^n_n \alpha^p_p \delta^p_{mn} = -\delta^p_{mn} = 0. \qquad (124)$$

All moduli are zero; the phenomenon cannot occur in a medium having a center of symmetry.

We shall now review the various possible cases of symmetry, whose association will enable us to find the possible piezoelectric properties in all crystalline systems.

9.1 CASE OF COMPLETE NONSYMMETRY

The tables of moduli for a case of complete nonsymmetry follow.

δ^1_{11}	δ^1_{12}	δ^1_{13}
δ^1_{12}	δ^1_{22}	δ^1_{23}
δ^1_{13}	δ^1_{23}	δ^1_{33}

δ^2_{11}	δ^2_{12}	δ^2_{13}
δ^2_{12}	δ^2_{22}	δ^2_{23}
δ^2_{13}	δ^2_{23}	δ^2_{33}

δ^3_{11}	δ^3_{12}	δ^3_{13}
δ^3_{12}	δ^3_{22}	δ^3_{23}
δ^3_{13}	δ^3_{23}	δ^3_{33}

9.2 CASE OF AN AXIS PASSING THROUGH ox_3, NONBINARY

The transformation matrix is

c	$-s$	
s	c	
		1

To effect the tensorial transformation

$$\bar{\delta}^k_{ij} = \sum_{lmn} \alpha^l_i \alpha^m_j \alpha^n_k \delta^n_{lm}$$

and write the relation

$$\delta^k_{ij} = \sum_{lmn} \alpha^l_i \alpha^m_j \alpha^n_k \delta^n_{lm}, \qquad (125)$$

we shall calculate all the possible terms equal to $\alpha^l_i \alpha^m_j$ by letting

$$\gamma^s_r = \alpha^l_i \alpha^m_j,$$

where r is a number designating the product ij, and s is a number designating the product lm.

As i, j, l, and m vary from 1 to 3, r and s vary from 1 to 9; then transformation (123) may be written

$$\delta^k_r = \sum_s \gamma^s_r \alpha^r_k \delta^n_s. \qquad (126)$$

This transformation can be effected by a matrix product as shown previously and (126) may be written in the form

$$(\delta) = (\gamma)(\delta)(\alpha^T). \qquad (127)$$

The terms of tensor δ^k_{ij} may be arranged in any order, provided matrix γ corresponds. In the matrix adopted here, lines 21, 31, and 32 are omitted as they give only equations already written. Consequently, the corresponding lines may be omitted in matrix γ, reducing it to 9 lines and 9 columns.

	1	2	3
11	δ^1_{11}	δ^2_{11}	δ^3_{11}
12	δ^1_{12}	δ^2_{12}	δ^3_{12}
22	δ^1_{22}	δ^2_{22}	δ^3_{22}
13	δ^1_{13}	δ^2_{13}	δ^3_{13}
23	δ^1_{23}	δ^2_{23}	δ^3_{23}
33	δ^1_{33}	δ^2_{33}	δ^3_{33}

To take account of the terms represented by the omitted lines 21, 31, and 32, matrix γ columns 12 and 21 are replaced by a column designated 12, which contains the sum of the terms of 12 and 21. Similar treatment is provided for 13 and 31, and for 23 and 32. The following arrangement is thus obtained for (127).

$$\begin{array}{c|c|c|c|} & 1 & 2 & 3 \\ \hline 11 & a & b & \gamma \\ \hline 12 & d & e & f \\ \hline 22 & g & h & i \\ \hline 13 & k & l & m \\ \hline 23 & n & \omega & p \\ \hline 33 & q & r & t \\ \hline \end{array} = \begin{array}{|c|c|c|c|c|c|} \hline c^2 & -2cs & s^2 & & & \\ \hline cs & c^2-s^2 & -c & & & \\ \hline s^2 & 2cs & c^2 & & & \\ \hline & & & c & -s & \\ \hline & & & s & c & \\ \hline & & & & & 1 \\ \hline \end{array} \times \begin{array}{|c|c|c|} \hline a & b & \gamma \\ \hline d & e & f \\ \hline g & h & i \\ \hline k & l & m \\ \hline n & o & p \\ \hline q & r & t \\ \hline \end{array} \times \begin{array}{|c|c|c|} \hline c & s & \\ \hline -s & c & \\ \hline & & 1 \\ \hline \end{array}. \tag{128}$$

It is evident that the matrix products given above are of the form

$$\begin{array}{|c|} \hline A \\ \hline B \\ \hline \end{array} = \begin{array}{|c|c|} \hline c & 0 \\ \hline 0 & \alpha \\ \hline \end{array} \times \begin{array}{|c|} \hline A \\ \hline B \\ \hline \end{array} \times \begin{array}{|c|} \hline \alpha^T \\ \hline \end{array}, \tag{129}$$

which gives two groups of equations.

$$\left.\begin{aligned} (A) &= (C)(A)(\alpha^T) \\ (B) &= (\alpha)(B)(\alpha^T). \end{aligned}\right\} \tag{130}$$

The second group has already been obtained for the case of two-index tensors and gives the relations:

$$k=\omega, \qquad n=-l, \qquad t=t,$$

and

$$m=p=q=r=0.$$

Also

$$\delta^1_{13}=\delta^2_{23}, \quad \delta^2_{13}=-\delta^1_{23}, \quad \delta^3_{13}=\delta^3_{23}=\delta^1_{33}=\delta^2_{33}=0,$$

and

$$\delta^{3'}_{33}=\delta^3_{33}.$$

The equations of the first group must now be developed.

$$\begin{array}{|c|c|c|} \hline a & b & \gamma \\ \hline d & e & f \\ \hline g & h & i \\ \hline \end{array} = \begin{array}{|c|c|c|} \hline c^2 & -2cs & s^2 \\ \hline c\delta & c^2-s^2 & -cs \\ \hline \delta^2 & 2cs & c^2 \\ \hline \end{array} \times \begin{array}{|c|c|c|} \hline a & b & \gamma \\ \hline d & e & f \\ \hline g & h & i \\ \hline \end{array} \times \begin{array}{|c|c|c|} \hline c & \delta & \\ \hline -s & c & \\ \hline & & 1 \\ \hline \end{array}. \tag{131}$$

$$\begin{array}{|c|c|c|} \hline a & b & \gamma \\ \hline d & e & f \\ \hline g & h & i \\ \hline \end{array} = \begin{array}{|c|c|c|} \hline c^2a-2csd+s^2g & c^2b-2csxe+s^2h & c^2\gamma-2csf+s^2i \\ \hline csa+(c^2-\delta^2)d-csh & csb+(c^2-s^2)e-csh & cs\gamma+(c^2s^2)f-csi \\ \hline s^2a+2csd+c^2g & s^2b+2cse+c^2h & s^2\gamma+2csf+c^2i \\ \hline \end{array} \times \begin{array}{|c|c|c|} \hline c & s & \\ \hline -s & c & \\ \hline & & 1 \\ \hline \end{array}, \tag{132}$$

from which we derive:

$$\left.\begin{aligned}(a+g)(c-1)-s(b+h)&=0\\(b+h)(c-1)+s(a+g)&=0\\(i-\gamma)s-2fc&=0\\-(i\gamma)c-2fs&=0\\(a+e)(c-1)-(d+h)s&=0\\-(a+e)s-(d+h)(c-1)&=0.\end{aligned}\right\}\qquad(133)$$

Hence,

$$a=-g=-e,\qquad -b=+h=-d,$$
$$i=\gamma,\qquad \text{and}\qquad f=0,$$

which gives the following arrangement of the table of moduli.

δ_{11}^1	$-\delta_{22}^2$	δ_{13}^1
$-\delta_{22}^2$	$-\delta_{11}^1$	δ_{23}^1
δ_{13}^1	δ_{23}^1	

1

$-\delta_{22}^2$	$-\delta_{11}^1$	$-\delta_{23}^1$
$-\delta_{11}^1$	δ_{22}^2	δ_{13}^1
$-\delta_{23}^1$	δ_{13}^1	

2

δ_{11}^3		
	δ_{11}^3	
		δ_{33}^3

3

(134)

9.3 CASE OF TERNARY AXIS, PARALLEL TO DIAGONAL OF CUBE WHOSE SIDES ARE PARALLEL TO AXES OF COORDINATES

In the case of a ternary axis that is parallel to the diagonal of a cube, the sides of which are parallel to the axes of coordinates, the relation (125) becomes

$$\delta_{ij}^k=\alpha_i^{i+1}\alpha_j^{j+1}\alpha_k^{k+1}\delta_{(i+1)(j+1)}^{(k+1)}.\qquad(135)$$

The three coefficients of α being equal to unity, the relation is

$$\delta_{ij}^k=\delta_{(i+1)(j+1)}^{(k+1)}.$$

The terms of tables 2 and 3, given below, are derived from the terms of table 1 by circular permutation.

δ_{11}^1	δ_{12}^1	δ_{13}^1
δ_{12}^1	δ_{22}^1	δ_{23}^1
δ_{13}^1	δ_{23}^1	δ_{33}^1

1

δ_{33}^1	δ_{13}^1	δ_{23}^1
δ_{13}^1	δ_{11}^1	δ_{12}^1
δ_{23}^1	δ_{12}^1	δ_{22}^1

2

δ_{22}^1	δ_{23}^1	δ_{12}^1
δ_{23}^1	δ_{33}^1	δ_{13}^1
δ_{12}^1	δ_{13}^1	δ_{11}^1

3

9.4 ALTERNATE PLANE

The transformation matrix for the alternate plane is

	1	2	3
1		-1	
2	1		
3			-1

.

The relations (124) have only one term in their right-hand members, the index of which is derived by a circular permutation of the indices of the left-hand member.

For certain terms, such as δ_{11}^1, we find that $\delta_{11}^1=-\delta_{22}^2$ and $\delta_{22}^2=+\delta_{11}^1$, whence $\delta_{11}^1=\delta_{22}^2=0$. By acting in this manner on all terms, the following tables are obtained.

	1	2	3
1	0	0	δ_{13}^1
2	0	0	δ_{23}^1
3	δ_{13}^1	δ_{23}^1	

1

1	2	3
0	0	δ_{23}^1
0	0	$-\delta_{13}^1$
δ_{23}^1	$-\delta_{13}^1$	

2

1	2	3
δ_{11}^3	δ_{12}^3	0
δ_{21}^3	$-\delta_{11}^3$	0
0	0	0

3

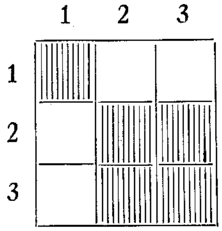
9.5 SYMMETRY WITH RESPECT TO BINARY AXES

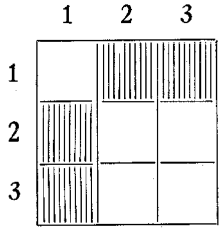
Symmetry with respect to the binary axes manifests itself by the suppression of certain moduli. If we assume the binary axis L_1^2, the transformation matrix and (124) become

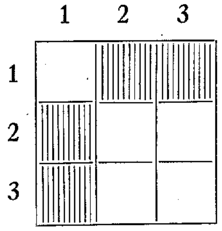
$$\begin{array}{c|c|c|c|} & 1 & 2 & 3 \\ \hline 1 & 1 & & \\ \hline 2 & & -1 & \\ \hline 3 & & & -1 \\ \hline \end{array}, \qquad (131)$$

$\delta^{k}_{ij} = \pm\delta^{k}_{ij}$. Every time the product $\alpha^{i}_{i}\alpha^{j}_{j}\alpha^{l}_{k} = -1$, the modulus cannot exist.

A grid may be formed by superposing 3 tables having numbered boxes of which those corresponding to a + sign have been cut out. The following arrangement is obtained for axis L_1^2.

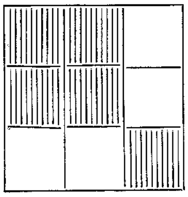

1

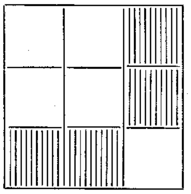

2

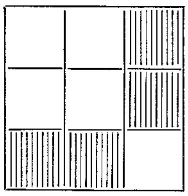

3

The transformation matrix and grids for axis $L_3^{(2)}$ are as follows.

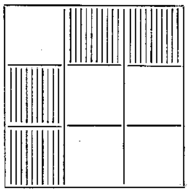
$$\begin{array}{|c|c|c|} \hline -1 & & \\ \hline & -1 & \\ \hline & & 1 \\ \hline \end{array}$$

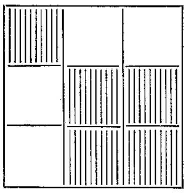


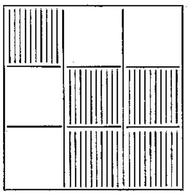


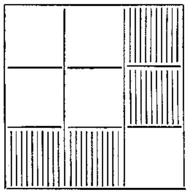


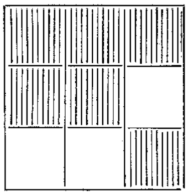
The grids corresponding to symmetries with respect to the planes are complementary to the preceding ones.

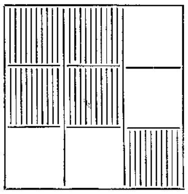

Plane P_1


Plane P_1


Plane P_1


Plane P_3


Plane P_3


Plane P_3

With these four tables of moduli and grids, it is easy to determine the piezoelectric properties of all crystalline groups by the moduli corresponding to the relation between the deformations and the fields.

We can represent, in a similar way, the relation between dielectric polarizations and stresses in a solid.

9.6 Expression of Stresses in Material

Consider a material in a state of stress and also in equilibrium. The stress may be represented by a small parallelepiped isolated from the rest of the material and maintained in the same condition of constraint by applying forces to its faces.

Thus, in Fig. 5, on face $BCDE$, a force FA is applied at the center of F and, in general, is oriented obliquely with respect to the normal of the face.

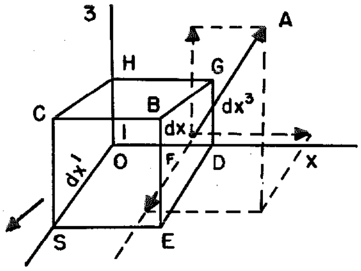


Fig. 5.

This force, relative to one face whose normal is directed along ox^2, is numbered $\vec{F}_2$ and has three components designated F_{12}, F_{22}, and F_{32}. These components are proportional to the areas of the surfaces to which they are applied, and their quotients by these areas represent effectively the stress condition at the point considered.

This quotient is represented by T.

$$T_{ij}=\frac{F_{ij}}{d\sigma j}. \tag{136}$$

On the face $GHIS$, a force is exerted that is equal and opposite in sign to FA.

If we designate the area $d\sigma$ as $+$ when the normal out of the cube through that surface is oriented in the positive direction of the axes, and designate the opposite case as $-$, the stresses T_{ij} are expressed in the same manner on two opposite faces and, therefore, nine quantities T_{ij} will be sufficient to represent the stresses at a point which is the center of the cube, when the dimensions of the cube are decreased.

There are really only six distinct quantities T_{ij}, i.e.,

$$T_{ij}=T_{ji} \tag{137}$$

for, if the system is in equilibrium, there is no resultant couple. The forces acting on the faces, projected on AB and OC, Fig. 6, are $T_{12}dx^1dx^3$, and their moments with respect to the axis of the cube parallel to ox^3 are

$$T=T_{12}dx^1dx^3dx^2=T_{12}d\sigma.$$

Similarly, the forces applied to the faces projected on CA and BC have a resultant moment

$$T=T_{21}d\sigma.$$

As the system is in equilibrium,

$$T_{12}=T_{21}.$$

These nine components constitute a tensor. If u^1 is the displacement of straight line OA after deformation, that of straight line BC is $u^1_1+du^1_1$, and the work of deformation effected by the forces oriented along axis 1 is

$$dw_1=\bar{F}_1du^1$$

since

$$F_1=T_{1j}d\sigma^j$$

and

$$dw_1=T_{ij}du^1d\sigma^j.$$

For the whole volume, the work is

$$dw=T_{ij}du^1d\sigma^j. \tag{138}$$

We shall represent all quantities in rectangular coordinates, and the coefficients of the transformation matrixes α are the directing cosines of the second system of axes with respect to the first. We know that under such conditions, the areas of the surfaces are transformed in accordance with the values of their normals. We thus have a right to consider the areas as being once contravariant; the displacements du, homogeneous to lengths, are also once contravariant; therefore the stresses are twice covariant.

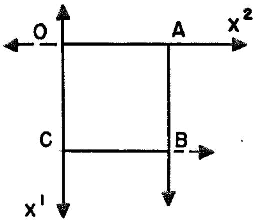


Fig. 6.

The piezoelectric relation can be written as follows.

In a constrained body, there appears a dielectric polarization that is a linear function of the deformations. The dielectric polarization is related to the induction

$$\vec{b} = K_0\vec{h} + 4\pi\vec{P}, \tag{139}$$

$$P_i = \delta_i^{jk} T_{jk}. \tag{140}$$

Lippmann showed by thermodynamic analogy the identity of one of the moduli δ_1^{11} with the modulus previously described. The same reasoning may be extended to all moduli, the δ_1^{jk} and the δ_{jk}^{i} are really the covariant and contravariant components of one tensor, components which are numerically equal in trirectangular axes.

We can easily see that the general relations (139) and (140) can assume the shape given originally by Curie for one particular case.

If we assume a slab of quartz cut perpendicularly to the electric axis and subjected to a compressional stress in a direction normal to the force, we have

$$P_1 = \delta_1^{11} T_{11}, \tag{141}$$

and if at the same time a potential difference V is applied between the electrodes covering the slab, we have

$$b_1 = K h_1 + 4\pi\delta_1^{11} T_{11}. \tag{142}$$

But $b = 4\pi\sigma$, where σ designates the surface density of the charges, hence

$$\sigma = \frac{K}{4\pi}\frac{v}{e} + \delta_1^{11} T_{11}, \tag{143}$$

and if the capacitor has an area S,

$$Q = \sigma s = \frac{KS}{4\pi e} V + \delta_1^{11} F_1, \tag{144}$$

$$Q = CV + \delta F, \tag{145}$$

which is the relation given initially by Curie.

10. Elasticity of Solids, Hooke's Law, Tensorial Moduli With Four Indices

Hooke's Law states that there exists a linear relation between the stresses and the strains in an elastic body.

Any particular stress T_{ij} caused in an anisotropic body by any particular strain T_{kl}, results in a modulus defined by

$$T_{ij} = C_{ij}^{kl} \delta_{kl}. \tag{146}$$

If the body is completely anisotropic, all the constraints are linear functions of all the strains, and (146) then represents a system of nine equations, each having nine terms in its right-hand member. Thus there are 81 moduli C_{ij}^{kl} forming the components of a tensor of the fourth order, twice covariant, and twice contravariant.

If we know the components of the tensor at point M, we can calculate the components at point $\bar{M}$, whose coordinates are derived from those of point M by means of the linear transformation matrix α and the general relation

$$\overline{C_{mn}^{pq}} = \alpha_m^i \alpha_n^j \beta_k^p \beta_p^q C_{ij}^{kl}, \tag{147}$$

or

$$C_{mn}^{pq} = \sum_{ijkl} \alpha_m^1 \alpha_n^j \alpha_p^k \alpha_q^l C_{ij}^{kl}. \tag{148}$$

Actually, there are not 81 independent moduli even in a completely anisotropic body.

As $\delta_{kl} = \delta_{lk}$ and $T_{ij} = T_{ji}$, we derive from (146) and for particular values of i, j, k, and l

$$C_{ij}^{kl} = C_{ji}^{kl} = C_{ij}^{lk} = C_{ji}^{lk}. \tag{149}$$

As the strain potential energy, which is invariant, can be determined by

$$dw = \sum_{ij} T_{ij} S_{ij},$$

we deduce

$$dw = \sum_{(ij)(kl)} C_{ij}^{kl} S_{ij} S_k^l. \tag{150}$$

In the latter sum, we can interchange the groups (ij) and (kl) and obtain

$$C_{ij}^{kl} = C_{kl}^{ij},$$

hence

$$C_{ij}^{kl} = C_{ji}^{kl} = C_{ij}^{kl} = C_{ji}^{kl} = C_{kl}^{ij} = C_{kl}^{ji} = C_{lk}^{ji} = C_{lk}^{ij} = C_{lk}^{ji}.$$

There remain, finally, only 21 independent moduli. There are actually only six components S_{ij} for the strain, and all the terms form a homogeneous second-degree polynomial with respect to these six quantities and, thus, include $6(6+1)/2 = 21$ independent coefficients.

For these, it is customary to use a notation having a reduced number of indices, but we prefer to keep the four distinct indices to facilitate tensorial transformations and circular permutations, which are fundamental in expressing certain properties of symmetry.

We shall represent the moduli in tables having six lines and six columns.

To calculate the relation between moduli where there is a nonbinary symmetry axis ox^3, we can use a method similar to that for the three-index moduli; by letting $\gamma_r^s = \alpha_m^i \alpha_n^j$ and $\gamma_t^u = \alpha_p^k \alpha_q^l$, (148) becomes

$$C_r^s = \sum_{su} \gamma_r^s \gamma_t^u C_s^u, \tag{151}$$

which can be calculated by a matrix product

$$(C) = (\gamma)(C)(\gamma^T). \tag{152}$$

Developing and designating momentarily the moduli a, b, etc., (their indices correspond to the numbers at the heads of the lines and columns) we have

Designating matrixes corresponding to one fourth of each of the above matrixes by A, B, C, D, E, the operations are symbolized by the relation

$$\begin{array}{|c|c|}\hline A & B \\ \hline B^T & C \\ \hline \end{array}$$

$$= \begin{array}{|c|c|}\hline D & O \\ \hline O & E \\ \hline \end{array} \times \begin{array}{|c|c|}\hline A & B \\ \hline B^T & C \\ \hline \end{array} \times \begin{array}{|c|c|}\hline D^T & \\ \hline & E^T \\ \hline \end{array},$$

which gives four groups of relations: $(A) = (D)(A)(D^T)$, $(B) = (D)(B)(E^T)$, $(C) = (E)(C)(E^T)$, and $(B)^T = (E)(B)^T(D^T)$. The fourth relation gives the same equations as the second.

But the fourth group is similar to (110), which enabled us to represent the properties of the two-index tensor, and the third group is similar to (131) of the three-index tensor; we derive from it

$$P = \delta, \quad Q = -q = 0, \quad r = t = 0, \quad u = u,$$

$$\begin{array}{c|c|c|c|c|c|c|} & 11 & 12 & 22 & 13 & 23 & 33 \\ \hline 11 & a & b & \gamma & d & e & f \\ \hline 12 & b & g & h & i & j & k \\ \hline 22 & & h & l & m & n & o \\ \hline 13 & d & i & m & p & q & r \\ \hline 23 & e & j & n & q & s & t \\ \hline 33 & f & k & o & r & t & u \\ \hline \end{array} = \begin{array}{|c|c|c|c|c|c|}\hline c^2 & -2cs & s^2 & & & \\ \hline cs & c^2-s^2 & -cs & & & \\ \hline s^2 & 2cs & c^2 & & & \\ \hline & & & c & s & \\ \hline & & & -s & c & \\ \hline & & & & & 1 \\ \hline \end{array} \times \begin{array}{|c|c|c|c|c|c|}\hline a & b & \gamma & d & e & f \\ \hline b & g & h & i & j & k \\ \hline \gamma & h & l & m & n & o \\ \hline d & i & m & p & q & r \\ \hline e & j & n & q & s & t \\ \hline f & k & o & r & t & u \\ \hline \end{array}$$

$$\times \begin{array}{|c|c|c|c|c|c|}\hline c^2 & cs & s^2 & & & \\ \hline -2cs & c^2-s^2 & +2cs & & & \\ \hline s^2 & -cs & c_2 & & & \\ \hline & & & c & -s & \\ \hline & & & s & c & \\ \hline & & & & & 1 \\ \hline \end{array}.$$

and

$$d = -m = -j, \quad n = -e = -i, \quad \sigma = j, \quad \text{and} \quad K = 0.$$

That is,

$$\begin{aligned} C_{11}^{13} &= -C_{22}^{13} = -C_{12}^{23}, \\ C_{22}^{23} &= -C_{11}^{23} = -C_{12}^{13}, \\ C_{22}^{33} &= C_{11}^{33} = C_{12}^{33} = 0. \end{aligned}$$

There remains only to calculate the equations:

$$\begin{array}{c|c|c|c|}
 & 11 & 12 & 22 \\
\hline
11 & a & b & \gamma \\
\hline
12 & b & g & h \\
\hline
22 & \gamma & h & l \\
\hline
\end{array}
=
\begin{array}{|c|c|c|}
\hline
c^2 & -2cs & s^2 \\
\hline
cs & c^2-s^2 & -cs \\
\hline
s^2 & 2cs & c^2 \\
\hline
\end{array}
\times
\begin{array}{|c|c|c|}
\hline
a & b & \gamma \\
\hline
b & g & h \\
\hline
\gamma & h & l \\
\hline
\end{array}
\times
\begin{array}{|c|c|c|}
\hline
c^2 & cs & s^2 \\
\hline
-2cs & c^2-s^2 & 2cs \\
\hline
s^2 & -cs & c^2 \\
\hline
\end{array}$$

$$=
\begin{array}{|c|c|c|}
\hline
c^2a-2bcs+\gamma s^2 & c^2b-2gc\delta+hs^2 & \gamma c^2-2hcs+ls^2 \\
\hline
acs+b(c^2-s^2)-\gamma cs & bcs+g(c^2-s^2)-hcs & \gamma c\delta+h(c^2-s^2)-lcs \\
\hline
s^2a+2bcs+\gamma c^2 & bs^2+2gcs+hc^2 & \gamma s^2+2hcs+lc^2 \\
\hline
\end{array}
\times
\begin{array}{|c|c|c|}
\hline
c^2 & cs & s^2 \\
\hline
-2cs & c^2-s^2 & 2cs \\
\hline
s^2 & -cs & c^2 \\
\hline
\end{array}.$$

Calculating the sum $a+\gamma$, as being the first and third terms of the first line, we see that all we have to do is to multiply the first line of the first matrix by the sum of the first and third columns of the second, i.e.,

$$\begin{array}{|c|}
\hline
1 \\
\hline
0 \\
\hline
1 \\
\hline
\end{array}.$$

We thus derive:

from $a+\gamma$,

$$0=\delta(l-a)-2(b+h)c;$$

from g,

$$(a-\gamma-2g)c\delta+(b-h)(c^2-\delta^2)=0;$$

from $b+h$,

$$0=-c(l-a)-2(b+h)\delta;$$

from $b-h$,

$$(a-\gamma-2g)(c^2-\delta^2)-(b-h)c\delta=0;$$

whence

$$l=a; \quad b=h=0; \quad g=\frac{a-\gamma}{2};$$

$$C_{11}^{11}=C_{22}^{22}; \quad C_{11}^{12}=C_{12}^{22}=0; \quad \text{and} \quad C_{12}^{12}=\frac{C_{11}^{11}-C_{11}^{22}}{2}.$$

The table of moduli follows.

	11	12	22	13	23	33
11	C_{11}^{11}		C_{11}^{22}	C_{11}^{13}	$-C_{22}^{23}$	C_{11}^{33}
12		$\frac{C_{11}^{11}-C_{11}^{22}}{2}$		$-C_{22}^{23}$	$-C_{11}^{13}$	
22			C_{11}^{11}	$-C_{11}^{13}$	C_{22}^{23}	C_{11}^{33}
13				C_{13}^{13}		
23					C_{13}^{13}	
33						C_{33}^{33}

There are only seven distinct moduli. This is the case of a ternary axis. The condensed table of moduli for a ternary axis is

	11	12	13	22	23	33
11	C_{11}^{11}	0	C_{11}^{13}	C_{11}^{22}	C_{11}^{23}	C_{11}^{33}
12		$\frac{C_{11}^{11}-C_{11}^{22}}{2}$	$-C_{11}^{23}$	0	$-C_{11}^{13}$	0
13			C_{13}^{13}	$-C_{11}^{13}$	0	0
22				C_{11}^{11}	$-C_{11}^{23}$	C_{11}^{33}
23					C_{13}^{13}	0
33						C_{33}^{33}

For a quaternary axis, the transformation matrix is

	-1	
$+1$		
		1

which gives the tables:

	11	12	13	22	23	33
11	C_{11}^{11}	C_{11}^{12}		C_{11}^{22}		C_{11}^{33}
12		C_{12}^{12}		$-C_{11}^{12}$		
13			C_{13}^{13}			
22				C_{11}^{11}		C_{11}^{33}
23					C_{13}^{13}	
33						C_{33}^{33}

Quaternary Axis

	11	12	13	22	23	33
11	C_{11}^{11}	C_{11}^{12}	C_{11}^{13}	C_{11}^{22}	C_{11}^{23}	C_{11}^{22}
12		C_{12}^{12}	C_{12}^{13}	C_{11}^{13}	C_{12}^{13}	C_{11}^{23}
13			C_{12}^{12}	C_{11}^{23}	C_{12}^{13}	C_{11}^{12}
22				C_{11}^{11}	C_{11}^{12}	C_{11}^{22}
23					C_{12}^{12}	C_{11}^{13}
33						C_{11}^{11}

Ternary Axis of the Cube

The table of moduli, in the case of symmetry about the diagonal of the cube, is derived from the table for complete nonsymmetry by circular permutation; the arrangement is obtained as follows.

When there exists an alternate plane of an even order, such as $\pi/2$, the symmetry about the binary axis L_3 makes the moduli having an odd number of indices vanish. The transformation matrix differs from that of the quaternary axis only in the sign of α_3^3, the table of moduli is the same as in the case of the quaternary axis, since α_3^3 is only involved by its square.

The grids corresponding to axis 1 and to the plane P_1 are the same as those of axis 3 and plane P_3, by virtue of the fact that the number of indices is even.

For symmetry about axis ox_1:

$$\alpha_1^1 = +1, \quad \alpha_2^2 = \alpha_3^3 - 1$$

for plane P_1,

$$\alpha_1^1 = -1, \quad \alpha_2^2 = \alpha_3^3 = +1.$$

If we consider the groups having an even number of "one" and if we express the symmetry, the relation will be of the form:

$$C_{ij}^{11} = \alpha_i^i \alpha_j^j C_{ij}^{11},$$

or

$$C_{ij}^{11} = (-1)^2 C_{ij}^{11}$$

with respect to the axis, and

$$C_{ij}^{11} = (+1)^2 C_{ij}^{11}$$

with respect to the plane.

If the group of indices contains index 1 only once,

$$C_{ij}^{lk} = (-1)^3 C_{ij}^{lk}$$

with respect to the axis, and

$$C_{ij}^{lk} = -1 C_{ij}^{lk}$$

with respect to the plane. This gives the same distribution of the signs + and −1.

	11	12	13	22	23	33
11	▨			▨	▨	▨
12		▨	▨			
13			▨			
22				▨	▨	▨
23					▨	▨
33						▨

L_1 Axis and Plane P_1

	11	12	13	22	23	33
11	▨	▨		▨		▨
12		▨		▨		▨
13			▨		▨	
22				▨		▨
23					▨	
33						▨

These tables give us information regarding the elastic properties of a material.

In the case of quartz, for instance, characterized by a ternary axis ox_3 and a binary axis ox_1, we obtain the following table:

	11	12	13	22	23	33
11	C_{11}^{11}			C_{11}^{22}	C_{11}^{23}	C_{11}^{33}
12		$\frac{C_{11}^{11} - C_{11}^{22}}{2}$	$-C_{11}^{23}$			
13			C_{13}^{13}			
22				C_{11}^{11}	$-C_{11}^{23}$	C_{11}^{33}
23					C_{13}^{13}	
33						C_{33}^{33}

We see, for instance, that a tensional stress oriented along the optical axis ox_3 can cause a strain in the directions 3 and δ_{33}, as well as in the directions 1 and 2, these two latter strains being equal, $\delta_{11}=\delta_{22}$.

Similarly, a stress of type T_{11} causes a strain δ_{11}, δ_{22}, and δ_{33}, and also a strain of type δ_{23}, etc.

We also see that a stress of type 12, caused for instance by a tensional torque whose moment is oriented along the electrical axis, causes a type-12 strain with a modulus

$$C_{12}^{12}=\frac{C_{11}^{11}-C_{11}^{22}}{2},$$

but also a type-11 strain, i.e., a change of length in the direction of axis 1, as would have been caused by a compression in the direction of this axis.

10.1 Case of Isotropic Materials

An isotropic material has an infinity of axes and of planes of symmetry; to reduce the number of moduli, we shall use the table for the ternary axis and superimpose the grids for binary axes L_1 and L_3; we have then found the arrangement and since there exists a symmetry about a ternary axis oriented as the cube diagonal, we must have equality between the moduli by circular permutation, which gives the tables shown below.

	11	12	13	22	23	33
11	C_{11}^{11}			C_{11}^{22}		C_{11}^{33}
12		$\frac{C_{11}^{11}-C_{12}^{11}}{2}$				
13			C_{13}^{13}			
22				C_{11}^{11}		C_{11}^{33}
23					C_{13}^{13}	
33						C_{33}^{33}

	11	12	13	22	23	33
11	C_{11}^{11}			C_{11}^{22}		C_{11}^{22}
12		$\frac{C_{11}^{11}-C_{11}^{22}}{2}$				
13			$\frac{C_{11}^{11}-C_{11}^{12}}{2}$			
22				C_{11}^{11}		C_{11}^{22}
23					$\frac{C_{11}^{11}-C_{11}^{22}}{2}$	
33						C_{11}^{11}

In the case of isotropic materials, there are only two distinct moduli, C_{11}^{11} and C_{11}^{22}. The first one corresponds to the proportionality relation between stress and strain in the direction of the stress. It has the same value along the three axes, and, therefore in all directions in space. It is Young's modulus E. The second one corresponds to a lengthening strain in a direction perpendicular to the tensional stress; it has the same value along the three axes, i.e., the lengthening is the same in all directions in space perpendicular to the stress. Poisson's coefficient gives the ratio between modulus C_{11}^{22} and Young's modulus. $C_{11}^{22}=\sigma E$. Finally, modulus C_{12}^{12} expresses the relation between a torque and the tension in a wire, it is proportional to Coulomb's modulus, but we see that it depends on moduli C_{11}^{11} and C_{11}^{22}, so that there are actually only two characteristic moduli for any isotropic material.

10.2 THERMAL EXPANSION OF SOLIDS

The change in length with temperature of a bar of length x may be expressed by the expansion binomial

$$x=x_0(1+A\theta),$$

where $\theta=t-t_0$, which can be written

$$\frac{x-x_0}{x_0}=A\theta.$$

But $(x-x_0)/x_0$ is the relative deformation in direction x, δ_{11} for instance. We can generalize and state that in an anisotropic body, a change in temperature may bring about a displacement of all points; these displacements are characterized by the tensor of the 3 and must produce a tensorial relation. The temperature change θ being a scalar $\delta_{ij}=A_{ij}x\theta$, i.e., a tensor of order zero, the expansion coefficient A forms a tensor with two covariant indices and six distinct components. The characteristic tables for the expansion coefficient for the various symmetries are the same as those for the specific inductive powers.

The tables of constants give only two temperature coefficients for certain crystalline substances; this constitutes a gap. In fact, by an appropriate choice of axes, we can always reduce the components of the tensor of the δ_{ij} to three components, $\bar{\delta}_{11}$, $\bar{\delta}_{22}$, $\bar{\delta}_{23}$, but in such a system there still exist three distinct components of the tensor of expansions.

10.3 THERMAL CONDUCTIVITY

Thermal conductivity may be defined by the amount of heat passing in one unit of time through an area σ^i under the action of a temperature gradient;

$$\frac{dq}{dt}=x_i^j\sigma^i\left(\frac{d\theta}{\partial x}\right)_j.$$

This gradient, in which θ has a zero variance and x is contravariant, is itself once covariant, and since dq/dt is invariant, the thermal conductivity must be measured in an anisotropic body by nine components out of which six are distinct.

The reductions caused by symmetry are the same as for the dielectric constant.

11. Piezomagnetism

By analogy with piezoelectricity, Voigt has studied the possibility of a deformation of matter placed in a magnetic field. If the phenomenon exists, its effect is very small and, to date, it does not seem to have been demonstrated clearly. It is interesting, however, to estimate which crystalline groups might exhibit this behavior.

We shall assume that the presence in the medium of a magnetic induction creates deformations. To find the relation between the moduli, we must first define the tensorial character of magnetic induction

Assume a charge e, moving at a velocity $\bar{v}$ in a magnetic field; the force applied to the mass is

$$\vec{f}=e\vec{v}A\vec{\beta},$$

and the work, which is an invariant, for a displacement dx of the charge is

$$d\mathcal{T}=\vec{f}\cdot\overrightarrow{dx},$$

$$d\mathcal{T}=(v^3\beta_3-v^2\beta_2)dx^1+(v^3\beta_1-v^1\beta_3)dx^2 +(v^1\beta_2-v^2\beta_1)dx^3.$$

This expression is of the form:

$$d\mathcal{T}=\beta_{ij}dx^iv^j,$$

with

$$\beta_{12}=\beta_3=-\beta_{21},$$
$$\beta_{23}=\beta_1=-\beta_{32},$$
$$\beta_{13}=-\beta_2=-\beta_{31}.$$

Magnetic induction appears here as a tensor of the second order, nonsymmetrical, twice covariant.

The phenomenon of piezomagnetism would be represented by moduli M^{kl}_{ij} such that

$$S_{ij}=M^{kl}_{ij}\beta_{kl}.$$

We see that

$$M^{kl}_{ij}=-M^{lk}_{ij}=M^{kl}_{ji}=-M^{pk}_{ji} \quad \text{and} \quad M^{ll}_{ij}=0.$$

There are 18 distinct moduli in the case of complete nonsymmetry. In this case, the table of moduli is

	11	12	13	21	22	23	31	32	33
11		M^{12}_{11}	M^{13}_{11}	$-M^{12}_{11}$		M^{11}_{23}	$-M^{13}_{11}$	$-M^{11}_{23}$	
12		M^{12}_{12}	M^{13}_{12}	$-M^{12}_{12}$		M^{23}_{12}	$-M^{13}_{12}$	$-M^{22}_{12}$	
13		M^{12}_{13}	M^{13}_{13}	$-M^{12}_{13}$		M^{23}_{13}	$-M^{13}_{13}$	$-M^{23}_{13}$	
21		M^{12}_{12}	M^{13}_{12}	$-M^{12}_{12}$		M^{23}_{12}	$-M^{23}_{12}$	$-M^{13}_{12}$	
22		M^{12}_{22}	M^{13}_{22}	$-M^{12}_{22}$		M^{23}_{22}	$-M^{13}_{22}$	$-M^{23}_{22}$	
23		M^{12}_{23}	M^{13}_{23}	$-M^{12}_{23}$		M^{23}_{23}	$-M^{13}_{23}$	$-M^{23}_{23}$	
31		M^{12}_{13}	M^{13}_{13}	$-M^{12}_{13}$		M^{23}_{13}	$-M^{13}_{13}$	$-M^{23}_{13}$	
32		M^{12}_{23}	M^{13}_{23}	$-M^{12}_{23}$		M^{23}_{23}	$-M^{13}_{23}$	$-M^{23}_{23}$	
33		M^{12}_{33}	M^{13}_{33}	$-M^{12}_{33}$		M^{23}_{33}	$-M^{13}_{33}$	$-M^{23}_{33}$	

11.1 REDUCTION OF MODULI FOR A NONBINARY AXIS

As in the case of elasticity, we group the indices in pairs and set the coefficients of a matrix γ so that

$$\gamma^r_s=\alpha^m_i\alpha^n_j.$$

The relations involved for the particular case of symmetry are written

$$(M)=(\gamma)(M)(\gamma^T).$$

The table of components of tensor M consists of nine lines and nine columns, but, because of nonsymmetry with respect to the upper indices, we need consider only three columns. Similarly, the symmetry of the lower indices, $\delta_{ij}=\delta_{ji}$, make necessary consideration of only six lines. This reduces matrix γ to only six lines and six columns, obtained by adding the terms of columns 1.2 +2.1, etc.In matrix (γ^T), we need consider only three lines and three columns. The equations are written

$$\begin{array}{|c|c|c|}\hline a & b & \gamma \\ \hline d & e & f \\ \hline g & h & i \\ \hline k & l & m \\ \hline n & \omega & p \\ \hline q & r & \sigma \\ \hline \end{array} = \begin{array}{|c|c|c|c|c|c|}\hline c^2 & -2cs & s^2 & & & \\ \hline +cs & c^2-s^2 & -cs & & & \\ \hline s^2 & +2cs & c^2 & & & \\ \hline & & & c & -s & \\ \hline & & & s & c & \\ \hline & & & & & 1 \\ \hline \end{array} \times \begin{array}{|c|c|c|}\hline a & b & \gamma \\ \hline d & e & f \\ \hline g & h & i \\ \hline k & l & m \\ \hline n & \omega & p \\ \hline q & r & \sigma \\ \hline \end{array} \times \begin{array}{|c|c|c|}\hline 1 & & \\ \hline & c & s \\ \hline & -s & c \\ \hline \end{array},$$

$$\begin{array}{|c|c|c|}\hline ac^2-2dcs+gs^2 & bc^2-2ecs+hs^2 & \gamma c^2-2fcs+is^2 \\ \hline (a-g)cs+d(c^2-s^2) & (b-h)cse(c^2-s^2) & (\gamma-i)cs+f(c^2-s^2) \\ \hline as^2+2dcs+gc^2 & bs^2+2ecs+gc^2 & (\gamma s^2)+2fcs+ic^2 \\ \hline kc-ns & lc-\omega s & mc-ps \\ \hline ks+nc & ls+\omega c & ms+pc \\ \hline q & r & \sigma \\ \hline \end{array} \times \begin{array}{|c|c|c|}\hline 1 & & \\ \hline & c & s \\ \hline & -s & c \\ \hline \end{array}.$$

Whence the equations:

$$\begin{array}{r|r|r}
(g-a)s-2dc=0 & (b+h)(c-1)-(\gamma+i)s=0 & (\gamma-e)(c-1)+s(h-f)=0 \\
-(g-a)c-2ds=0 & (b+h)s+(\gamma+i)(c-1)=0 & (\gamma-e)s-(h-f)(c-1)=0 \\
k(c-1)-ns=0 & +(p-l)c+(\omega+m)s=0 & \\
ks+n(c-1)=0 & r(c-1)-\sigma s=0 & \\
q=q & rs+\sigma(c-1)=0 &
\end{array}$$

$$a=g,\quad b=h=-f,\quad \gamma=e=-i,\quad p=l,\quad \omega=-m,\quad h=f,\quad d=k=n=p=\sigma=0.$$

The tables corresponding to the symmetry of the various axes are:

	12	13	23
11	M_{11}^{12}	M_{11}^{13}	M_{11}^{23}
12		M_{11}^{23}	$-M_{11}^{13}$
22	M_{11}^{12}	$-M_{11}^{13}$	$-M_{11}^{23}$
13		M_{13}^{13}	M_{13}^{23}
23		$-M_{13}^{23}$	M_{13}^{13}
33	M_{33}^{12}		

Ternary Axis

	12	13	23
11	M_{11}^{12}	M_{11}^{13}	M_{11}^{23}
12	M_{12}^{12}	M_{12}^{13}	M_{12}^{23}
22	M_{22}^{12}	M_{22}^{13}	M_{22}^{23}
13	M_{13}^{12}	M_{13}^{13}	M_{13}^{23}
23	M_{23}^{13}	M_{23}^{13}	M_{23}^{23}
33	M_{33}^{12}	M_{33}^{13}	M_{33}^{23}

Nonsymmetry

M^{12}_{11}	$-M^{12}_{22}$	M^{12}_{33}
M^{12}_{12}	$-M^{12}_{23}$	M^{12}_{13}
M^{12}_{22}	$-M^{12}_{33}$	M^{12}_{11}
M^{12}_{13}	$-M^{12}_{12}$	M^{12}_{13}
M^{12}_{23}	$-M^{12}_{13}$	M^{12}_{12}
M^{12}_{33}	$-M^{12}_{11}$	M^{12}_{22}

Cube Diagonal

M^{12}_{11}		
M^{12}_{11}		
	M^{13}_{13}	
		M^{13}_{13}
M^{12}_{33}		

Alternate Plane π_2

The grids for planes P_1 and P_3 are the same as those for the perpendicular binary axes.

	12	13	23
11			▨
12	▨	▨	
22			▨
13	▨	▨	
23			▨
33			▨

	12	13	23
11	▨		
12	▨		
22	▨		
13		▨	▨
23		▨	▨
33	▨		

12. *Physical Properties of Bodies Crystallizing in Various Systems*

12.1 CRYSTALLINE NETWORK

A crystal is capable of cleavage in several directions; the faces appearing by cleavage are plane. A crystalline network is constituted by three systems of planes that limit the smallest possible volume that can be isolated. These planes are called *reticular planes*, the smallest elementary volume is the *mesh*, and the apexes are the *nodes*. The arrangement of atoms inside the mesh is the *pattern*. The crystal network, most generally, presents a certain symmetry. The symmetry of the network is generally greater than that of the crystalline medium.

When the crystalline medium has the symmetry of its network, it is said to be holohedral. When the symmetry of the crystal is lower than that of the network, we have a merihedrism.

This merihedrism is a hemihedrism, tetartohedrism, or ogdohedrism when the crystal symmetry is one half, one fourth, or one eighth that of the network.

The number of natural faces of the crystal is equal to the number of faces of the network in the case of holohedrism, of twice, four times, or eight times that number in the case of hemihedrism, tetartohedrism, and ogdohedrism.

12.2 CRYSTALLINE SYSTEMS

The 32 classes of symmetry in crystals are divided into seven systems according to the symmetry of the crystal networks.

1. Triclinic System. The network has no axis of symmetry.
2. Binary or Monoclinic System. The network has one binary axis.
3. Orthorhombic or Terbinary System. The network has three binary axes, rectangular two by two.
4. Rhombohedral or Ternary System. The network has one ternary axis.
5. Quadratic System. The network has one quaternary axis.
6. Senary or Hexagonal System. The network has one senary axis.
7. Cubic System. The network has four ternary axes, three quaternary axes, and six binary axes.

We shall review rapidly these various systems as well as their classes of symmetry, indicating for each class a particular form and the table of components of the various characteristic tensors. We shall begin with the least nonsymmetric system.

12.2.1 Triclinic System

12.2.1.1 Holohedrism C

This form has no symmetry axis but has one center of symmetry, as shown in Fig. 7. All odd-order tensors are zero; it has no pyroelectric properties.

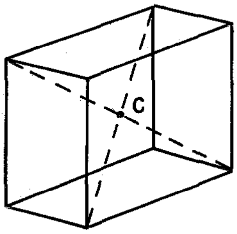


Fig. 7.

Dielectric Susceptance

x_1^1	x_1^2	x_1^3
x_2^1	x_2^2	x_2^3
x_3^1	x_3^2	x_3^3

9 Constants

Elasticity

	11	12	22	13	23	33
11	C_{11}^{11}	C_{11}^{12}	C_{11}^{22}	C_{11}^{13}	C_{11}^{23}	C_{11}^{33}
	12	C_{12}^{12}	C_{12}^{22}	C_{12}^{13}	C_{12}^{23}	C_{12}^{33}
		22	C_{22}^{22}	C_{22}^{13}	C_{22}^{23}	C_{22}^{33}
			13	C_{13}^{13}	C_{13}^{23}	C_{13}^{33}
				23	C_{23}^{23}	C_{23}^{33}
					33	C_{33}^{33}

21 Constants

Examples: albite, disthene.

12.2.1.2 Hemihedrism

This form has no symmetry. All even- or odd-order tensors are complete, the material may exhibit pyroelectricity or piezoelectricity.

Pyroelectricity

A_1
A_2
A_3

Piezoelectricity

δ_{11}^1	δ_{12}^1	δ_{13}^1
δ_{12}^1	δ_{22}^1	δ_{23}^1
δ_{13}^1	δ_{23}^1	δ_{33}^1

1

δ_{11}^2	δ_{12}^2	δ_{13}^2
δ_{12}^2	δ_{22}^2	δ_{23}^2
δ_{13}^2	δ_{23}^2	δ_{33}^2

2

δ_{11}^3	δ_{12}^3	δ_{13}^3
δ_{12}^3	δ_{22}^3	δ_{23}^3
δ_{13}^3	δ_{23}^3	δ_{33}^3

3

The tensors for dielectric susceptance and for elasticity are the same as in the preceding case. Example: axinite.

12.2.2 Monoclinic or Clinorhombohedral System

12.2.2.1 Holohedrism L^2PC

The holohedral form is a straight prism on a parallelogram as a base, as may be seen in Fig. 8.

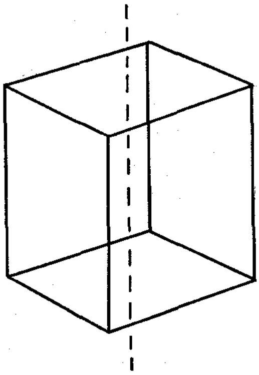


Fig. 8.

Dielectric Susceptance

x_1^1	x_1^2	
x_2^1	x_2^2	
		x_3^3

5 Moduli, Double-Refracted
Biaxial Crystals

Elasticity

C_{11}^{11}		C_{11}^{22}			C_{11}^{33}
	C_{12}^{12}				
		C_{22}^{22}			C_{22}^{33}
			C_{13}^{13}		
					C_{23}^{33}

8 Distinct Moduli

Neither pyroelectric nor piezoelectric properties exist.

12.2.2.2 Holoaxial Enantiomorphous Hemihedrism

This form has one binary axis $L_3^{(2)}$. The structure is shown in Fig. 9.

Pyroelectricity

π_3

1 Modulus

Dielectric Susceptance

x_1^1	x_1^2	
x_2^1	x_2^2	
		x_3^3

5 Moduli

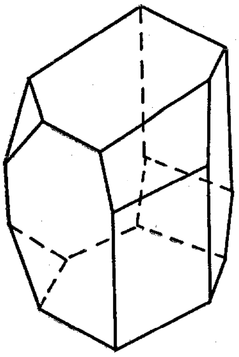


Fig. 9.

Elasticity

C_{11}^{11}	C_{11}^{12}	C_{11}^{22}			C_{11}^{33}
	C_{12}^{12}	C_{12}^{22}			
		C_{22}^{22}			C_{22}^{33}
			C_{13}^{13}	C_{13}^{23}	
					C_{33}^{33}

11 Moduli

Pyroelectricity

	1	2	3
1	δ_{11}^1	δ_{12}^1	
2	δ_{12}^1	δ_{22}^1	
3			δ_{33}^1

1	2	3
		δ_{13}^2
		δ_{23}^2
δ_{13}^2	δ_{23}^2	

1	2	3
		δ_{13}^3
		δ_{23}^3
δ_{13}^3	δ_{23}^3	

13 Moduli

Examples: tartaric acid, sugar, lithium sulfate.

12.2.2.3 Antihemihedrism

The crystalline form is shown in Fig. 10.

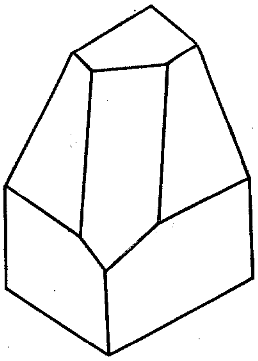


Fig. 10.

Pyroelectricity

π_2
π_3

2 Moduli

Dielectric Susceptance

	x_1^2	x_1^3
x_2^1		
x_3^1		

4 Moduli

One Plane P

	δ^1_{12}	δ^1_{13}
δ^1_{12}		
δ^1_{13}		

1

δ^2_{11}		
	δ^2_{22}	δ^2_{23}
	δ^2_{23}	δ^2_{33}

2

δ^3_{11}		
	δ^3_{22}	δ^3_{23}
	δ^3_{23}	δ^3_{33}

3

14 Moduli

Elasticity

11	12	22	13	23	33
C^{11}_{11}		C^{22}_{11}		C^{23}_{11}	C^{33}_{11}
	C^{12}_{12}		C^{13}_{12}		
		C^{22}_{22}		C^{23}_{22}	C^{33}_{22}
			C^{13}_{13}		
				C^{23}_{23}	C^{33}_{23}
					C^{33}_{33}

13 Moduli

Examples: organic compounds, copper sulfate.

12.2.3 *Orthorhombohedral System*

12.2.3.1 *Holohedrism*

This form has three perpendicular binary axes and one center as may be seen in Fig. 11. The original solid is an orthorhombohedral prism. All pyroelectric and piezoelectric moduli are zero.

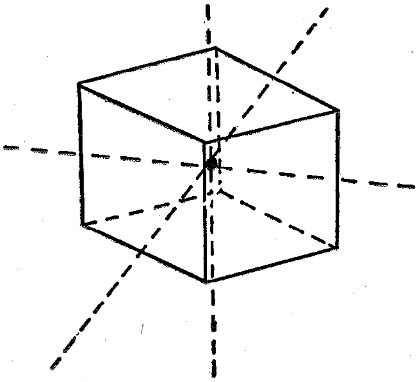


Fig. 11.

Dielectric Susceptance

x^1_1		
	x^2_2	
		x^3_3

3 Moduli

No Double Refraction

Elasticity

C^{11}_{11}		C^{22}_{11}			C^{33}_{11}
	C^{12}_{12}				
		C^{22}_{22}			C^{33}_{22}
			C^{12}_{13}		
					C^{33}_{23}

8 Moduli

Examples: sulphur, tridymite, aragonite.

12.2.3.2 *Holoaxial Hemihedrism* $L^{(2)}_3 L^{(2)}_1$

This form has no center and no pyroelectric properties. It is shown in Fig. 12.

Dielectric Susceptance

x^1_1		
	x^2_2	
		x^3_3

3 Moduli

	1	2	3
1			
2			δ^1_{23}
3		δ^1_{23}	

1

		δ^2_{13}
δ^2_{13}		

2

	δ^3_{12}	
δ^3_{12}		

3

3 Moduli of Type δ^1_{23}

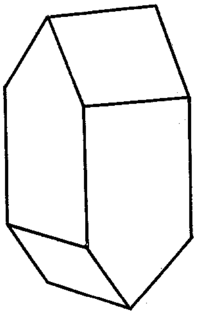


Fig. 12.

The electric field oriented in the direction of one of the crystallographic axes causes an angular strain in the perpendicular plane. For elasticity, it has the same moduli as holohedrism.

Examples: Rochelle salt, epsomite (magnesium sulphate).

12.2.3.3 Antihemihedrism $L_3^{(2)}P_1$

The crystalline form is shown in Fig. 13.

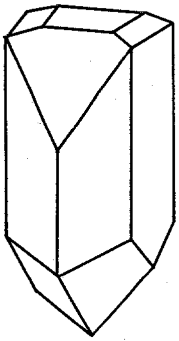


Fig. 13.

Pyroelectricity

π_3

1 modulus

Susceptance

	x_1^2	
x_2^1		

Piezoelectricity

		δ_{13}^1
δ_{13}^1		

1

		δ_{23}^2
	δ_{23}^2	

2

δ_{11}^3		
	δ_{22}^3	
		δ_{33}^3

3

5 Moduli

Examples: calamine, topaz, picric acid.

12.2.4 Rhombohedral System

12.2.4.1 Holohedrism $CL_3^3P_1$

The network is a rhombohedron as is evident from Fig. 14.

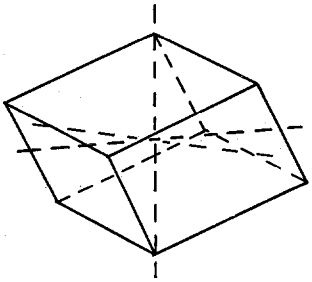


Fig. 14.

Dielectric Susceptance

	x_1^2	
$-x_1^2$		

Elasticity

C_{11}^{11}		C_{11}^{22}		$-C_{22}^{23}$	C_{11}^{33}
	$\frac{C_{11}^{11}-C_{11}^{22}}{2}$		$-C_{22}^{23}$		
		C_{11}^{11}		C_{22}^{23}	C_{11}^{33}
			C_{13}^{13}		
				C_{13}^{13}	
					C_{33}^{33}

7 Distinct Moduli

Examples: calcite, sodium nitrate, corindon, etc.

12.2.4.2 *Holoaxial Hemihedrism* $L_3^{(3)}L_1^{(2)}$

This form displays no pyroelectric properties. It is shown in Figs. 15 and 16. For elasticity, it has the same moduli as holohedrism.

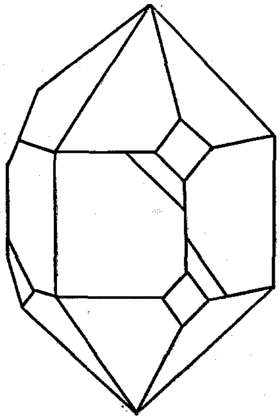


Fig. 15.

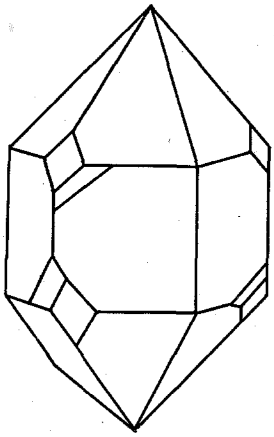


Fig. 16.

Dielectric Susceptance

x_1^1		
	x_1^1	
		x_3^3

3 Moduli

Piezoelectricity

	1	2	3
1	δ_{11}^1		
2		$-\delta_{11}^1$	δ_{23}^1
3		δ_{23}^1	

1

	1	2	3
1		$-\delta_{11}^1$	$-\delta_{23}^1$
2	$-\delta_{11}^1$		
3	$-\delta_{23}^1$		

2

	1	2	3
1			
2			
3			

3

2 Distinct Moduli

Example: quartz.

12.2.4.3 Antihemidrism $L_3^3P_1$

Pyroelectricity

π_3

Dielectric Susceptance

x_1^1		
	x_1^1	
		x_3^3

Piezoelectricity

	δ_{22}^2	δ_{13}^1
δ_{22}^2		
δ_{13}^1		

1

$-\delta_{22}^2$		
	δ_{22}^2	δ_{13}^1
	δ_{13}^1	

2

δ_{11}^3		
	δ_{11}^3	
		δ_{33}^3

3

For piezoelectricity, the electric field oriented along axis ox^3 causes a strain of the type δ_{33} and two equal strains in the perpendicular directions $\delta_{11}=\delta_{22}$, i.e., a tourmaline disk cut perpendicularly to axis ox^3 is compressed in the electric field at the same time it increases in width, remaining circular.

This form has the same moduli of elasticity as quartz. The structure is shown in Figs. 17, 18, and 19.

Examples: tourmaline, carborundum, potassium bromate.

12.2.4.4 Tetartohedrism L_3^3

The form of this crystal is shown in Fig. 20.

Pyroelectricity

π_3

Dielectric Susceptance

x_1^1	x_1^2	
$-x_1^2$	x_1^1	
		x_3^3

Piezoelectricity

δ_{11}^1	δ_{22}^2	δ_{13}^1
δ_{22}^2	$-\delta_{11}^1$	δ_{23}^1
δ_{13}^1	δ_{23}^1	

1

$-\delta_{22}^2$	$-\delta_{11}^1$	$-\delta_{23}^1$
$-\delta_{11}^1$	δ_{22}^2	δ_{13}^1
$-\delta_{23}^1$	δ_{13}^1	

2

δ_{11}^3		
	δ_{11}^3	
		δ_{33}^3

3

6 Moduli

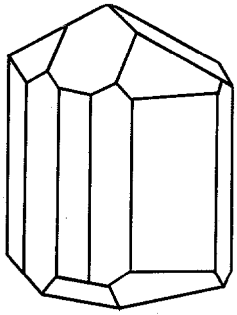


Fig. 17.

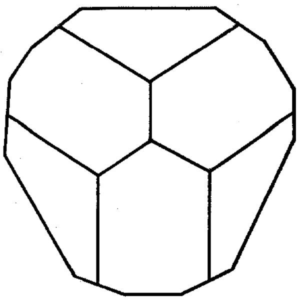


Fig. 18.

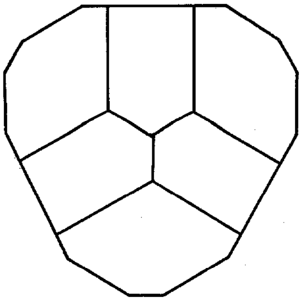


Fig. 19.

Elasticity

C^{11}_{11}		C^{22}_{11}	C^{13}_{11}	$-C^{23}_{22}$	C^{33}_{11}
	$\frac{C^{11}_{11}-C^{22}_{11}}{2}$		$-C^{23}_{22}$	$-C^{13}_{11}$	
		C^{11}_{11}	$-C^{13}_{11}$	C^{23}_{22}	C^{33}_{11}
			C^{13}_{13}		
				C^{13}_{13}	
					C^{33}_{33}

7 Moduli

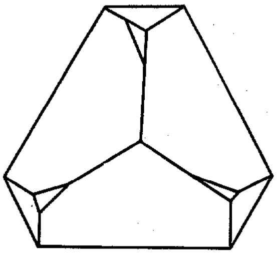


Fig. 20.

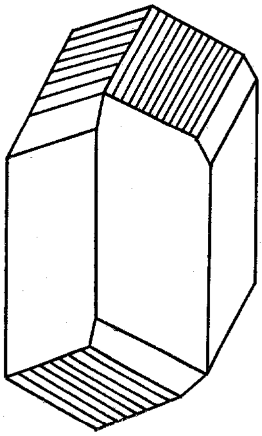


Fig. 21.

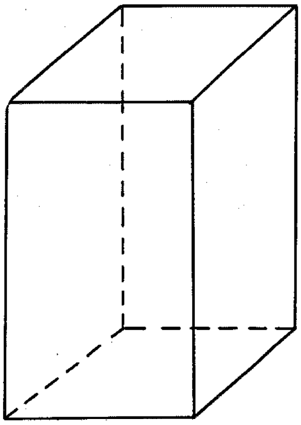


Fig. 22.

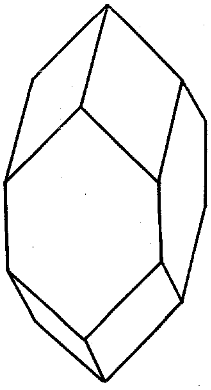


Fig. 23.

Example: sodium periodate.

12.2.4.5 Centered Hemihedrism L^3_3C

This form has no pyroelectric and no piezoelectric properties. Susceptance and elasticity are as for tetartohedrism. Fig. 21 shows the form of the crystal.

Example: dolomie.

12.2.5 Quadratic System

12.2.5.1 Holohedrism $L^{(4)}_3L^{(2)}_1P_1C$

The network consists of quadratic prisms. No pyroelectric and no piezoelectric properties exist. The crystalline form is shown in Fig. 22 and in Fig. 23.

Susceptance

x^1_1		
	x^1_1	
		x^3_3

Elasticity

	11	12	22	13	23	33
11	C_{11}^{11}		C_{11}^{22}			C_{11}^{33}
12		C_{12}^{12}				
22			C_{11}^{11}			C_{11}^{33}
13				C_{13}^{13}		
23						
33						C_{33}^{33}

7 Moduli

In the case of the quadratic system, the reduction of the moduli are the same as in the case of the hexagonal system, the 90-degree rotation gives the same reductions as a rotation of 120 degrees, and the additional 180-degree rotation is the same in both cases. The only difference is in the modulus of elasticity C_{12}^{12}.

Rotation about axis 3, of any angle, gives the relation

$$C^2\delta^2 \left| C_{12}^{12} - \frac{C_{11}^{11} - C_{11}^{22}}{2} \right| = 0.$$

But, here, for 90 or 180 degrees, $C_3 = 0$; the relation always holds and the condition

$$C_{12}^{12} = \frac{(C_{11}^{11} - C_{11}^{12})}{2}$$

is not necessary.

Example: calomel.

12.2.5.2 *Holoaxial Hemihedrism* $L_3^{(4)}L_1^2$

This form has no pyroelectricity. For dielectric susceptance, it has the same moduli as holohedrism. The structure of the crystal is shown in Fig. 24.

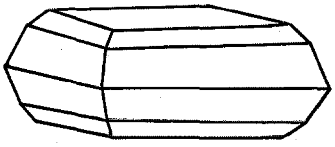


Fig. 24.

Piezoelectricity

		δ_{23}^{1}
	δ_{23}^{1}	

1

		$-\delta_{23}^{1}$
$-\delta_{23}^{1}$		

2

3

Elasticity

C_{11}^{11}		C_{11}^{22}			C_{11}^{33}
	C_{12}^{12}				
		C_{11}^{11}			C_{11}^{33}
			C_{13}^{13}		
				C_{13}^{13}	
					C_{33}^{33}

8 Moduli

Example: nickel sulphate.

12.2.5.3 *Antihemihedrism* $L_3^{(4)}P_1$

Pyroelectricity

π_3

Dielectric Susceptance

	x_1^2	
$-x_1^2$		

Piezoelectricity

		δ_{13}^{1}
δ_{13}^{1}		

		δ_{13}^{1}
	δ_{13}^{1}	

δ_{11}^{3}		
	δ_{11}^{3}	
		δ_{33}^{3}

This form has the same moduli of elasticity as the preceding class. Fig. 25 shows the crystalline form.

Example: chalcopyrite.

12.2.5.4 *Parahemihedrism* $\Lambda^4 C$

This form has no pyroelectricity and no piezoelectricity. It has the same dielectric susceptances as holohedrism and the same elasticity moduli as holoaxial hemihedrism. The form may be seen in Fig. 26.

Example: schellite.

12.2.5.5 *Tetartohedrism With Quaternary Axis* $L_3^{(4)}$

The shape of this crystal is shown in Fig. 27.

Pyroelectricity

π_3

Dielectric Susceptance

x_1^1		
	x_1^1	
		x_3^3

Piezoelectricity

		δ_{13}^1
		δ_{23}^1
δ_{13}^1	δ_{23}^1	

1

		δ_{23}^1
		δ_{13}^1
$-\delta_{23}^1$	δ_{13}^1	

2

δ_{11}^3		
	δ_{11}^3	
		δ_{33}^3

3

For elasticity, this form has the same moduli as hemihedrism.

Example: lead molybdate.

12.2.5.6 *Sphenohedral Antihemihedrism* $L_3^{(2)} L_1^{(2)} \bar{\omega}_2$

This form has no pyroelectric properties. It has the same elasticity moduli as hemihedrism. The form is shown in Fig. 28.

Dielectric Susceptance

x_1^1		
	x_2^2	
		x_3^3

Piezoelectricity

		δ_{23}^1
	δ_{23}^1	

1

		δ_{23}^1
δ_{23}^1		

2

	δ_{12}^3	
δ_{12}^3		

3

Example: mercuric cyanide, tetraethyl ammonium iodide in sphenoidal form.

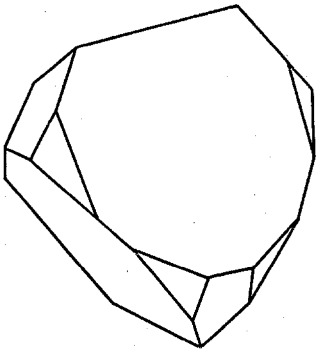


Fig. 25.

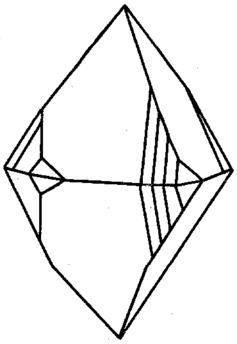


Fig. 26.

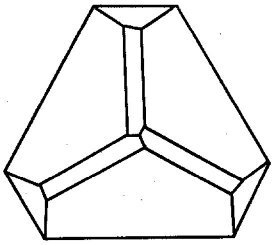


Fig. 27.

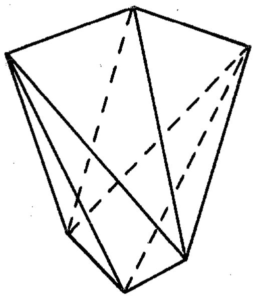


Fig. 28.

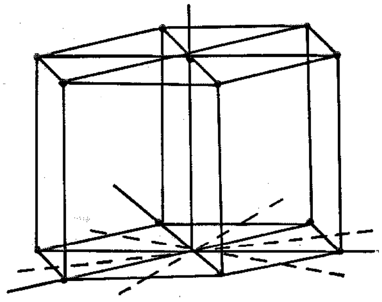


Fig. 30.

12.2.5.7 *Sphenohedral Tetartohedrism* $L_3^2\tilde{\omega}_2$

No crystals exist in this class. It would have the shape shown in Fig. 29.

12.2.6 *Hexagonal System*

12.2.6.1 *Holohedrism* $L_3^6L_1P_1C$

In primitive form, it is a multiple mesh in the shape of a regular hexagonal prism as shown in Fig. 30. It has no pyroelectric and no piezoelectric properties.

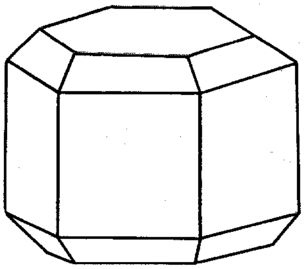


Fig. 31.

Dielectric Susceptance

x_1^1		
	x_1^1	
		x_3^3

Elasticity

	11	12	22	13	23	33
11	C_{11}^{11}		C_{11}^{22}			C_{11}^{33}
12		$\frac{C_{11}^{11}-C_{11}^{22}}{2}$				
22			C_{11}^{11}			C_{11}^{33}
13				C_{13}^{13}		
23					C_{13}^{13}	
33						C_{33}^{33}

5 Distinct Moduli

Example: magnesium. (See Fig. 31.)

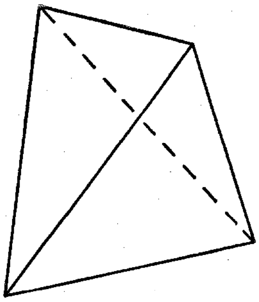


Fig. 29.

12.2.6.2. *Holoaxial Hemihedrism* $L_3^{(6)}L_1^{(2)}$

This form has no pyroelectric properties. It has the same dielectric susceptances and the same elasticity moduli as holohedrism.

Piezoelectricity

		δ_{23}^{1}
	δ_{23}^{1}	

1

		$-\delta_{23}^{1}$
$-\delta_{23}^{1}$		

2

3

Example: β quartz above 575 degrees centigrade.

12.2.6.3 Parahemihedrism With Ternary Axis $L_3^{(6)}P_3C$

This form has no pyroelectric and no piezoelectric properties. It has the same dielectric susceptances and same elasticity moduli as holohedrism. The form is shown in Fig. 32.

Example: apatite.

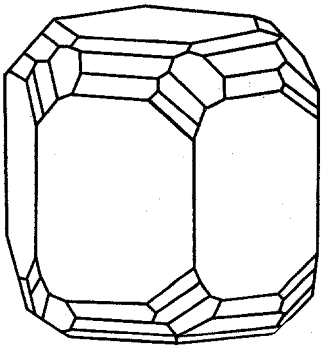


Fig. 32.

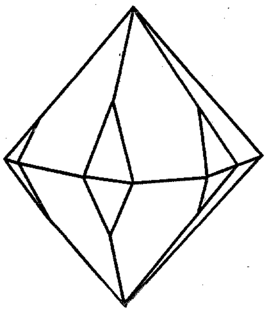


Fig. 33.

12.2.6.4 Antihemihedrism $\Lambda^{(6)}P_1$

Its dielectric susceptance and elasticity have the same moduli as holohedrism. The shape is shown in Fig. 33.

Pyroelectricity

π_3

Piezoelectricity

	1	2	3
1			δ_{13}^{1}
2			
3	δ_{13}^{1}		

1

	1	2	3
1			
2			δ_{13}^{1}
3		δ_{13}^{1}	

2

	1	2	3
1	δ_{11}^{3}		
2		δ_{11}^{3}	
3			δ_{33}^{3}

3

Examples: zincite, wurzite, iodyrite.

12.2.6.5 Tetartohedrism $L_3^{(6)}$

For elasticity, this form has the same moduli as holohedrism.

Pyroelectricity

π_3

Dielectric Susceptance

x_1^1	x_1^2	
$-x_1^2$	x_1^1	
		x_3^3

Piezoelectricity

		δ_{13}^1
		δ_{23}^1
δ_{13}^1	δ_{23}^1	

		$-\delta_{23}^1$
		δ_{13}^1
$-\delta_{23}^1$	δ_{13}^1	

δ_{11}^3		
	δ_{11}^3	
		δ_{33}^3

Examples: double lithium, potassium sulphate. (See Fig. 34.)

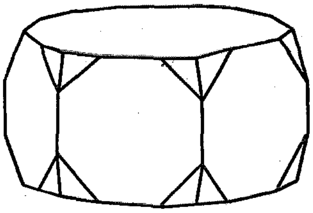


Fig. 34.

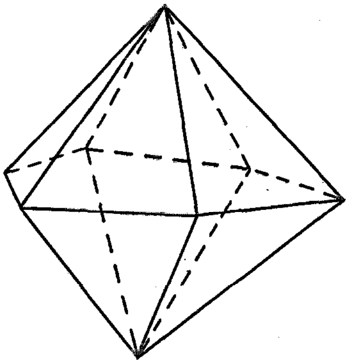


Fig. 35.

12.2.6.6 Trigonohedral Antihemihedrism

$L_{(3)}^3 L_1^{(2)} P_3$

This form has no pyroelectric properties. The shape is given in Fig. 35.

Dielectric Susceptance

x_1^1		
	x_1^1	
		x_3^3

Piezoelectricity

	11	12	13
11	δ_{11}^1		
12		$-\delta_{11}^1$	
13			

1

	$-\delta_{11}^1$	
$-\delta_{11}^1$		

2

3

Elasticity

C_{11}^{11}		C_{11}^{22}		$-C_{22}^{23}$	C_{11}^{33}
	$\frac{C_{11}^{11}-C_{11}^{22}}{2}$		$-C_{22}^{13}$		
		C_{11}^{11}		C_{22}^{23}	C_{11}^{23}
			C_{13}^{13}		
				C_{13}^{13}	
					C_{33}^{33}

Example: benitoite.

12.2.6.7 Trigonohedral Antitetartohedrism $\Lambda^3 P_3$

Pyroelectricity

π_3

Dielectric Susceptance

x_1^1	x_1^2	
$-x_1^2$	x_1^1	
		x_3^3

Piezoelectricity

δ_{11}^1	δ_{22}^2	
δ_{22}^2	$-\delta_{11}^1$	

1

$-\delta_{22}^2$	$-\delta_{11}^1$	
$-\delta_{11}^1$	δ_{22}^2	

2

3

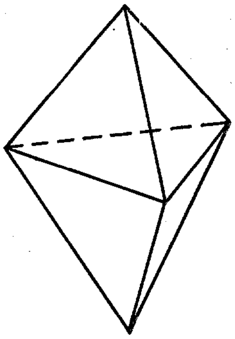


Fig. 36.

Example: This form has very rare symmetry and exists, perhaps, as silver phosphate. (See Fig. 36.)

12.2.7 Cubic System

12.2.7.1 Holohedrism $L_3^{(4)}L'^{3}_{3}L_1P_1C$

This form has no pyroelectric and no piezoelectric properties. The elasticity and dielectric susceptance moduli are the same as in an isotropic body. The crystalline form is shown in Fig. 37.

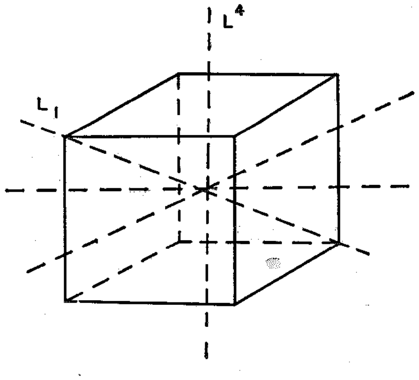


Fig. 37.

Dielectric Susceptance

x_1^1		
	x_1^1	
		x_1^1

Elasticity

C_{11}^{11}		C_{11}^{22}			C_{11}^{23}
	$\frac{C_{11}^{11}-C_{11}^{22}}{2}$				
		C_{11}^{11}			C_{11}^{23}
			C_{13}^{13}		
				C_{13}^{13}	
					C_{33}^{33}

Elasticity

11	12	22	13	23	33
C_{11}^{11}		C_{11}^{22}			C_{11}^{22}
	$\frac{C_{11}^{11}-C_{11}^{22}}{2}$				
		C_{11}^{11}			C_{11}^{22}
			$\frac{C_{11}^{11}-C_{11}^{22}}{2}$		
				$\frac{C_{11}^{11}-C_{11}^{22}}{2}$	
					C_{11}^{11}

Example: diamond, rock salt.

12.2.7.2 *Holoaxial Hemihedrism* $L_3^{(4)}L'^3L_1$

This form, shown in Fig. 38, has only one dielectric susceptance. It has no pyroelectric or piezoelectric properties.

Piezoelectricity

		δ_{23}^{1}
	δ_{23}^{1}	

1

		δ_{23}^{1}
	δ_{23}^{1}	

2

	δ_{23}^{1}	
δ_{23}^{1}		

3

Elasticity

C_{11}^{11}		C_{11}^{22}			C_{11}^{22}
	C_{12}^{12}				
		C_{11}^{11}			C_{11}^{22}
			C_{12}^{12}		
				C_{12}^{12}	
					C_{11}^{11}

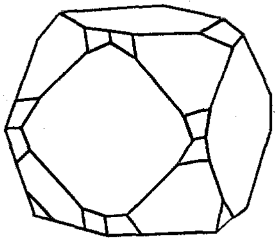


Fig. 38.

Example: sylvine.

12.2.7.3 *Parahemihedrism* $L_3^2L_3'P_3C$

This form has no piezoelectric properties. It has the same elastic constants as holoaxial hemihedrism. The structure is given in Fig. 39.

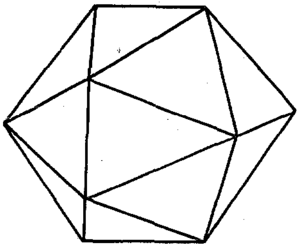


Fig. 39.

Examples: pyrite, cobaltine, alums.

12.2.7.4 *Antihemihedrism* $L_3^{(2)}L_3'\ \bar{\omega}_2P_1$

This form, the structure of which is given in Fig. 40, has no pyroelectric or piezoelectric properties. For elasticity, it has the same moduli as holo-axial hemihedrism.

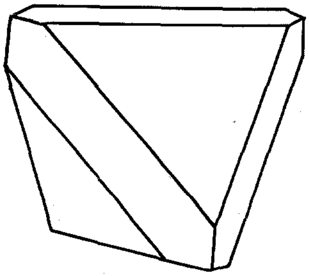


Fig. 40.

Example: blende.

12.2.7.5 *Tetartohedrism* $L_3^2L_3'L_1$

This form has no pyroelectric or piezoelectric properties. For elasticity, it has the same moduli as holoaxial hemihedrism. Its structure may be seen in Fig. 41.

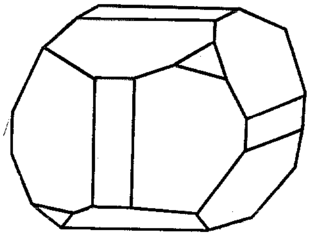


Fig. 41.

Example: sodium chlorate.

13. *Conclusions*

From these examples, it is evident that the particular symmetry of various crystalline systems indicate certain phenomena that are likely to take place in the crystallized material.

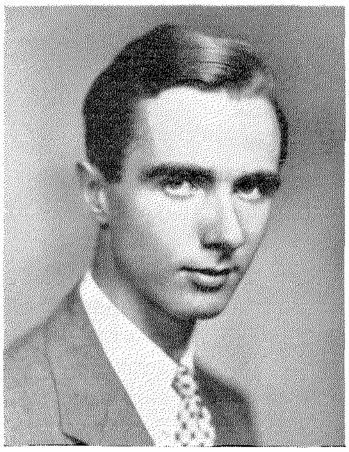
Tensor and matrix calculus permit the moduli linking cause and effect for all physical phenomena to be indicated for any crystalline system.

If a physical effect represented by a tensor of the order p is caused by the action represented by a tensor of the order q, the properties of the matrix act through a particular modulus that is a tensor of the order $p+q$.

The analytical transformations implied by this tensorial characteristic enable the relation between these moduli, as determined by the particular symmetry of the system, to be written. For instance, all tensorial moduli of odd order are zero if the crystalline substance has a center of symmetry.

It should be noted that the conclusions derived from this analysis indicate only the possibility of the existence of the various moduli. Such existence may not be stated with certainty. Only knowledge of the intimate structure of a crystal will allow its physical properties to be predicted.

Contributors to This Issue



Bernard T. Ellis, Jr.

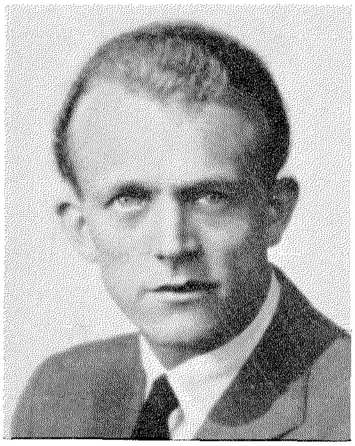
Bernard T. Ellis, Jr. was born in St. Thomas, V. I., on June 29, 1921. He received the B.S. degree in electrical engineering at Newark College of Engineering in 1942. Since 1942, he has been engaged in development and product engineering of radio communication transmitters with Federal Telephone and Radio Corporation.

He holds associate memberships in the American Institute of Electrical Engineers and in the Institute of Radio Engineers.

* * *

Fernand Mortiaux was born in Antwerp on March 3, 1906. He finished his studies in electromechanical engineering in 1929 at the University of Brussels.

Mr. Mortiaux entered the Institute National Belge de Radiodiffusion in 1930 and has been active in all phases of broadcasting. He was named chief engineer in 1935 and in 1946 became Administrateur-Directeur Général of the technical department.



Fernand Mortiaux

Mr. Mortiaux is a reserve officer; he is a member of various technical committees, including the Committee on Television.

* * *

Albert Shadowitz was born on May 5, 1915 at Brooklyn, N. Y. He received the B.E.E. degree from Polytechnic Institute of Brooklyn, in 1935. After attending Harvard University during 1936 and 1937, he studied at the University of California from 1937 to 1940.

He served as an engineering employee at Aberdeen Proving Grounds from 1941 to 1943. In 1943, he joined Federal Telecommunication Laboratories as an engineer.

Mr. Shadowitz is an Associate Member of the Institute of Radio Engineers.

* * *

Martin Silver graduated from New York University in 1941 with a Bachelor of Engineering degree. He then came to Federal Telecommunication Laboratories where he worked on communication-type jamming equipment. At the termination of the war, he was assigned to the development of frequency-modulation broadcast transmitters, monitors, links, and associated equipment.

Mr. Silver is an Associate Member of the Institute of Radio Engineers.

* * *

Marcel Tournier was born in Paris on March 23, 1888. In 1909, he received a degree in engineering from Ecole de Physique et de Chimie Industrielles in Paris, to which he returned in 1911, after obtaining the grade of Licencié ès Sciences, as an assistant in general chemistry. He became chief of general physics laboratory studies in 1914.

He collaborated with Professor P. Langevin in Paris and at the Laboratoire du Centre d'Etudes de la Marine at Toulon. They developed a complete equipment for submarine detection incorporating piezoelectric apparatus. In 1928, he joined Les Laboratoires, Le Matériel Téléphonique as head of the piezoelectric quartz department.

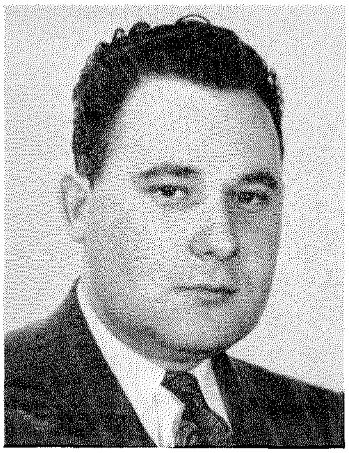
Albert Shadowitz

In 1937, Mr. Tournier was named electrotechnical professor at the Ecole de Physique et de Chimie, and in 1941 conference lecturer of the radio section at the Ecole Supérieure d'Electricité. He is also serving as head of the research division of the Office National d'Etudes et de Recherches Aéronautiques.

He was awarded the Medal of the Société d'Encouragement pour l'Industrie Nationale for his early developments of the electronic organ. He is a Chevalier de la Légion d'Honneur and a member of the Société Française de Physique.

* * *

For the biography and photograph of E. P. G. Wright, see Volume 24, Number 1, page 128.



Martin Silver

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[2] Cable service. [3] International and Marine Radiotelegraph services.
[4] Cable and Radiotelegraph services. [5] Radiotelegraph service.

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